VOLUME 26 JULY, 1938 NUMBER 7

PROCEEDINGS
of
The Institute of Radio Engineers

Application Blank for Associate Membership on Page IX

Institute of Radio Engineers Forthcoming Meetings

PACIFIC COAST CONVENTION
Portland, Oregon
August 10 and 11, 1938

CLEVELAND SECTION
September 22, 1938

DETROIT SECTION
September 16, 1938

LOS ANGELES SECTION
September 20, 1938

NEW YORK MEETING
October 5, 1938

PHILADELPHIA SECTION
September 1, 1938

PITTSBURGH SECTION
September 20, 1938

WASHINGTON SECTION
September 12, 1938

Pacific Coast Convention

A Pacific Coast Convention will be held in Portland, Oregon, on August 10 and 11, with headquarters at the Multnomah Hotel. This meeting is being held in conjunction with the convention of the American Institute of Electrical Engineers which will run from August 9 to 12. The program of papers to be presented at our meetings is given below and summaries of a number of the papers, arranged alphabetically by the names of the authors, follow it.

Wednesday, August 10

10:00 A.M.—12:00 Noon

"Radio Communication and Equipment Development in the Field of National Forest Protection," by H. K. Lawson, United States Forest Service. (Demonstration.)

"A Study of the Penetration of Radio Waves into Reinforced Concrete Buildings," by A. V. Eastman, H. M. Swarm, and William Harrold, University of Washington.

"Some Developments in Preselectors and Electric Tuners," by H. F. Elliott, consultant. (Demonstration.)

"An Automatic Light-Signaling Device for the Photoelectric Control of Fog Signals," by Marcus O'Day, Reed College.

2:00 P.M.—5:00 P.M.

(Broadcast Symposium)

"Selecting the Best Method of Applying Negative Feedback to Various Types of Transmitters," by L. S. Bookwalter, Radio Stations KOIN and KALE.

"Broadcast Antennas," by N. D. Webster, McClatchy Broadcasting Company.

"Doherty Amplifier Design," by James Middlebrooks, Columbia Broadcasting System.

"International-Broadcast-License Engineering," by E. G. Pack, Radio Station KSL.

"Studio and Transmitter Housing Construction," by R. V. Howard, Radio Station KSFO. (Demonstration.)

8:15—9:30 P.M.

"Electronic Applications in the Motion-Picture Industry," by G. S. Mitchell, Research Council of the Academy of Motion Picture Arts and Sciences. (Demonstration.)

Thursday, August 11

9:00 A.M.—12:00 Noon

(Joint Session with American Institute of Electrical Engineers)

"An Image-Transmission System Suitable for Telephone Signals," by A. M. Skellett, Bell Telephone Laboratories, Inc.

"Devices for Controlling Amplitude Characteristics of Telephone Signals," by A. C. Norwine, Bell Telephone Laboratories, Inc.

"Loud-Speaker Considerations in Feed-Back Amplifiers," by H. S. Knowles, Jensen Manufacturing Company.

"Some Developments and Problems of Demountable-Tube Design," by C. V. Litton, Litton Engineering Laboratories.

"Practical Application of Facsimile Broadcasting," by H. C. Singleton, Radio Stations KGW and KEX.

2:00 P.M.—5:00 P.M.

"Some Applications of Negative Feedback with Particular Reference to Laboratory Equipment," by F. E. Terman, R. R. Buss, and F. C. Cahill, Stanford University.

"An Improved Audio-Frequency Oscillator," by W. R. Hewlett. (Demonstration.)

"Practical Aspects of Wide-Band Television-Amplifier Design," by F. A. Everest, Oregon State College.

"An Impedance Bridge for Measurements at a Frequency of 200 Megacycles," by W. R. Hill, University of California.

SUMMARIES

SELECTING THE BEST METHOD OF APPLYING NEGATIVE FEEDBACK TO VARIOUS TYPES OF TRANSMITTERS

L. S. Bookwalter
(Radio Stations KOIN and KALE)

This paper describes the advantages of using negative feedback over only the audio-frequency stages in high-level transmitters. This paper also discusses other problems encountered in practical application of inverse feedback such as the cause of instability in certain types of transmitters employing negative feedback when modulated over 100 per cent, a method of employing negative feedback in push-pull stages in transmitters and eliminating the usual poor results encountered, the different places in which phase shift occurs and methods of holding this phase shift to a minimum, methods of obtaining stability by reducing the feedback at frequencies at which it becomes positive, and the necessity of using audio-frequency transformers that are designed for feed-back application.

A STUDY OF THE PENETRATION OF RADIO WAVES INTO REINFORCED CONCRETE BUILDINGS

A. V. Eastman, H. M. Swarm, and William Harrold
(University of Washington)

A study was made of the relative amount of signal received inside reinforced concrete buildings at various frequencies from 2000 to 40,000 kilocycles. The effect of both horizontal and vertical antennas was studied. The purpose of the work was to determine the frequency most suitable for establishing communication with patrolmen or other individuals inside of buildings, when they are equipped with portable receiving sets.

SOME DEVELOPMENTS IN PRESELECTORS AND ELECTRIC TUNERS

H. F. Elliott
(Consultant)

Broadcast-receiver-design attention has recently turned to a means for increasing the convenience and utility of the broadcast receiver. The past year or two has seen the introduction of a wide variety of mechanical and electrical dials and push-button arrangements to facilitate tuning.

This paper describes two electric tuners and a preselector. The first of the two tuners has been in commercial use since the first of 1938 for push-button tuning in automobile receivers. Six buttons mounted in a small control head on the dash enable the driver to tune in six stations without distracting his attention from driving.

The second tuner to be described comprises a plurality of notched disks with provisions for clamping them upon a rotor which is connected with the variable condenser. Six (or eight) magnets actuate latches engaging the notches. Tuning may be adjusted by the lay operator without the help of a serviceman.

The preselector comprises a synchronous electric clock with conventional hands and face for indicating time. A twenty-four-hour hand within the mechanism contacts successively ninety-six small rods. A knob-and-finger index dial provides convenient means for setting the connectors for any desired series of programs.

PRACTICAL ASPECTS OF WIDE-BAND TELEVISION-AMPLIFIER DESIGN

F. A. Everest
(Oregon State College)

Most of the literature covering video-frequency amplifiers deals only with the theoretical circumstances surrounding the factors limiting the high- and low-frequency, steady-state, and transient response of resistance-capacitance-coupled amplifiers and methods of extending the useful response region. This paper offers an extension of this information and its application to practical video-frequency amplifiers.

AN IMPROVED AUDIO-FREQUENCY OSCILLATOR

W. R. Hewlett

This paper describes a new type of oscillator in which the frequency-determining element is a resistance-capacitance network with no coils. The oscillator operates by a combination of positive and negative feedback and will cover the frequency band 20 to 20,000 cycles in three ranges. The oscillator is characterized by good wave form, high frequency stability, and constant amplitude over all the ranges. There are no critical adjustments, and the construction can follow the standard methods used in building audio-frequency amplifiers. This new oscillator offers the possibility of an inexpensive substitute for the ordinary beat-frequency oscillator.

AN IMPEDANCE BRIDGE FOR MEASUREMENTS AT A FREQUENCY OF 200 MEGACYCLES

W. R. Hill
(University of California)

The necessity for measurements at frequencies in the neighborhood of 200 megacycles is cited. The difficulties encountered in the construction of measuring equipment for such frequencies are discussed. A method of design and construction of a bridge suitable for this application is described and a model of this bridge is exhibited.

STUDIO AND TRANSMITTER-HOUSING CONSTRUCTION

R. V. Howard
(Radio Station KSFO)

The problems encountered in the construction of the new transmitter plant and studios for Station KSFO are discussed with particular reference to building design. A motion picture showing various stages of this construction augments this paper. Reference is also made to problems encountered in studio construction and arrangement.

LOUD-SPEAKER CONSIDERATIONS IN FEED-BACK AMPLIFIERS

H. S. Knowles
(Jensen Manufacturing Company)

The performance of a loud speaker depends on the characteristics of the source feeding it. Systems employing feedback from the electrical, mechanical, or acoustical meshes of a loud speaker must be considered as a whole.

The driving-point impedances of typical speakers, including various acoustical loads, are derived, and experimentally determined values given. The effects of the source and feedback on the steady-state and transient response of some typical systems are given.

RADIO COMMUNICATION AND EQUIPMENT DEVELOPMENT IN THE FIELD OF NATIONAL FOREST PROTECTION

H. K. Lawson
(United States Forest Service)

An outline is given of the general communication needs in the protection and administration of the national forests. The application of radio in this service is discussed and the design factors that must be considered are covered. The establishment and functions of the United States Forest Service radio laboratory is described and the methods employed in attacking its problems are pointed out. A description of several pieces of equipment used in this service will be given and several types of high-frequency and ultra-high-frequency equipment will be displayed.

SOME DEVELOPMENTS AND PROBLEMS OF DEMOUNTABLE-TUBE DESIGN

C. V. Litton
(Litton Engineering Laboratories)

The general problems and historical background of demountable-tube structures are discussed. A compact high-speed pumping unit is described, which is permanently attached to the tube. A tube design is described in which the cathode structures are renewable without use of special apparatus or techniques. The general application of the system is discussed from an operational viewpoint. A sample unit will be on display.

INTERNATIONAL-BROADCAST-LICENSE ENGINEERING

E. G. Pack
(Radio Station KSL)

The problem of collecting material and arranging it in proper form for application for an international broadcast license is reviewed. Unusual problems peculiar to this form of engineering are discussed.

PRACTICAL APPLICATION OF FACSIMILE BROADCASTING

H. C. Singleton
(Radio Stations KGW and KEX)

This paper discribes the purposes and practical application of facsimile broadcasting. It also discusses the make-up of the material to be broadcast. A technical description of the RCA equipment to be installed at KGW is provided.

SOME APPLICATIONS OF NEGATIVE FEEDBACK WITH PARTICULAR REFERENCE TO LABORATORY EQUIPMENT

F. E. Terman, R. R. Buss, and F. C. Cahill
(Stanford University)

Various applications of negative feedback in laboratory amplifiers are described. These include a substitute for a vacuum-tube voltmeter in which an amplifier stabilized by negative feedback delivers its output to a thermocouple, a standard-gain amplifier in which known values of gain such as 10, 20, and 50 are stabilized by means of negative feedback and used to increase the sensitivity of vacuum-tube and feed-back voltmeters, tuned amplifiers including those in which the amplification and selectivity are independent of tube constants and tube voltages, and means for applying negative feedback to laboratory oscillators to reduce the distortion of the output wave for all load impedances from short circuit to open circuit.

The use of negative feedback to develop a stabilized negative resistance substantially independent of tubes and supply voltages is considered. The application of such a stabilized negative resistance in increasing the selectivity of tuned circuits, in improving the alternating-to-direct-current-impedance ratio in diode detectors, and in resistance-tuned circuits, is discussed.

A tuned amplifier is shown in which the selectivity can be varied by means of a volume control without affecting the gain at resonance. This circuit is capable of giving equivalent Q's for the maximum selective condition which are of the order of 10 to 20 times the Q of the actual tuned circuit employed. The value of such selective circuits in wave-analyzer use is pointed out and the design of an analyzer employing them is given.

BROADCAST ANTENNAS

N. D. Webster
(McClatchy Broadcasting Company)

This paper reviews common practice in vertical-antenna design and describes a new method of top loading. Field surveys and other material is presented. The advantages and characteristics of this system are discussed.

June Meeting of the Board of Directors

A special meeting of the Board of Directors was held in the Institute office on Wednesday, June 1, 1938. Those present were H. H. Beverage, acting chairman; F. W. Cunningham, Alan Hazeltine, C. M. Jansky, Jr., B. J. Thompson, H. M. Turner, and H. P. Westman, secretary. The meeting was devoted to a detailed consideration of the recommendations of the Constitution and Laws Committee in regard to a revision of the Institute Constitution. A revised document was approved and will be mailed to the membership for final vote shortly.

Committee Work

Admissions Committee

A meeting of the Admissions Committee was held in the Institute office on Wednesday, May 4, and attended by F. W. Cunningham, chairman; I. S. Coggeshall (representing E. R. Shute), Melville Eastham, R. A. Heising, and H. P. Westman, secretary. Two of three applications for transfer to Member and one for admission to that grade were approved.

Board of Editors

The Board of Editors met in the Institute office on Tuesday, April 19. Those present were Alfred N. Goldsmith, chairman; R. R. Batcher, R. S. Burnap (representing B. E. Shackelford), P. S. Carter, J. D. Crawford, advertising manager; Helen M. Stote, assistant editor; K. S. Van Dyke, and H. P. Westman, secretary.

A careful analysis of the existing Proceedings format and possibilities of improving the presentation of the material now published, together with the high probability of increased advertising income resulted in a recommendation to the Board of Directors that the

Proceedings, starting with the January, 1939, issue be published in a trim size of eight and one-half inches by eleven inches and an advertising page size of seven inches by ten inches. It is anticipated that this advertising page size will permit a substantial increase in revenue and publication of the equivalent of several hundred additional pages per year without the necessity of obtaining further funds directly from the membership. It was agreed that if this change was made, a new cover design, new paper stock, and a clearer type face also be recommended. Realizing the problems involved in making changes in any publication after many years of existence, the Board of Editors nevertheless felt that existing conditions made imperative that this modification of format be adopted.

Constitution and Laws Committee

On Wednesday, May 25, the Constitution and Laws Committee met in the Institute office and completed its revision of the Institute Constitution. Those present were H. M. Turner, chairman; B. J. Thompson, J. D. Crawford, assistant secretary, and H. P. Westman, secretary.

Convention Committee

On April 22 and May 3, meetings of the Convention Committee were held in the Institute office. The April meeting was attended by H. P. Westman, chairman and secretary; E. K. Cohan, J. D. Crawford, advertising manager, P. B. Harkins (representing J. R. Poppele), C. E. Horn, C. E. Scholz, Helen M. Stote, B. J. Thompson, A. F. Van Dyck, and William Wilson.

The May meeting was attended by H. P. Westman, chairman and secretary; Austin Bailey, E. K. Cohan, I. S. Coggeshall (representing E. R. Shute), P. B. Harkins (representing J. R. Poppele), C. W. Horn, C. E. Scholz, Helen M. Stote, and B. J. Thompson.

The Convention Papers Subcommittee met in the Institute office on April 21 and May 3, and William Wilson, chairman; J. D. Crawford, assistant secretary; A. F. Van Dyck, and H. P. Westman, secretary, attended both of these meetings.

These four meetings were held in order to make the arrangements for our Thirteenth Annual Convention.

Registration of Professional Engineers Committee

A meeting of the Committee on the Registration of Professional Engineers was held in the Institute office on May 6 and attended by L. M. Hull, chairman; L. C. F. Horle, E. L. Nelson, A. P. Richmond (guest), A. F. Van Dyck, and H. P. Westman, secretary. Mr. Rich-

mond is the assistant secretary of the American Institute of Civil Engineers and has been active in the field of professional engineering registration.

Another meeting of the committee was held on June 2 and attended by L. M. Hull, chairman; C. R. Beardsley (guest), A. F. Van Dyck, and J. D. Crawford, assistant secretary. At this meeting the committee had the benefit of the experience of Mr. Beardsley who has been very active in this work in the American Institute of Electrical Engineers.

Membership Committee

Two meetings of the Membership Committee were held in May in the Institute office. The first, on the 4th, was attended by C. E. Scholz, chairman; H. A. Chinn, I. S. Coggeshall, Coke Flannagan, H. C. Gawler, R. M. Heintz, L. G. Pacent, C. R. Rowe, and J. D. Crawford, assistant secretary.

The meeting on the 12th was attended by C. E. Scholz, chairman; H. A. Chinn, I. S. Coggeshall, E. D. Cook, J. D. Crawford, assistant secretary; F. W. Cunningham, H. C. Gawler, H. J. Vennes, and H. P. Westman, secretary.

Both these meetings were devoted to a discussion of portions of the Institute Constitution which concern membership. Specific recommendations were prepared for submission to the Constitution and Laws Committee.

American Standards Association

Technical Committee on Radio Receivers

The Technical Committee on Radio Receivers, operating under the Sectional Committee on Radio of the American Standards Association, met in the Institute office on May 11. Those present were G. L. Beers, chairman; D. E. Foster, C. J. Franks, J. W. Fulmer, F. A. Polkinghorn, and H. P. Westman, secretary. The meeting was devoted to a discussion of some documents submitted by the International Electrotechnical Commission. Comments were prepared for submission to a conference which will consider the material for possible international standardization.

A number of manufacturing standards were recommended for adoption by the American Standards Association.

Technical Committee on Vacuum Tubes

On Tuesday, May 24, the Technical Committee on Vacuum tubes of the Sectional Committee on Radio, met in the Institute office. F. R.

Lack, chairman; R. S. Burnap (representing D. F. Schmit), N. P. Case, J. D. Crawford (guest), and H. P. Westman, secretary, were present.

This committee also commented on material submitted by the International Electrotechnical Commission for international standardization.

Institute Meetings

Atlanta

On April 21 a meeting of the Atlanta Section, attended by thirty-six, was held at the Georgia School of Technology with C. F. Daugherty chairman, presiding.

I. H. Gerks, professor of electrical engineering at the Georgia School of Technology, presented a paper on "The ABC of Television." The subject was first considered from a financial viewpoint and the great cost of transmitting and receiving equipment necessary to permit general use of the system was emphasized. The need for the adoption of standards on the technical operation of the equipment was pointed out as a factor holding up the introduction of television for the public. There was then presented a technical description of television systems as developed to date and the paper was concluded with a brief discussion of ultra-high-frequency-wave propagation. A general discussion followed.

Buffalo-Niagara

G. C. Crom, Jr., chairman, presided at the April 20 meeting of the Buffalo-Niagara Section held at the University of Buffalo and attended by twenty.

R. F. Field, engineer for the General Radio Company, presented a paper on "The Schering Bridge." The historical background of the bridge was first reviewed. It was pointed out that as in the case of the Wheatstone bridge, the actual inventors' names were not given to the device. The ideas embodied in the Schering bridge were described in 1914 by Phillips Thomas whereas Schering presented his work in 1920.

The bridge equations were developed about two factors D and Q. $D=(R/X)$ was called the dissipation factor and $Q=(X/R)$ was called the storage factor. There followed a survey of the errors caused by residual impedances in the bridge arms. A discussion of the terminal impedances of the transformer and various methods of grounding the bridge were given. The various uses of the bridge and its limitations were described.

Chicago

On April 19, the Chicago Section met jointly with the local section of the American Institute of Electrical Engineers and the electrical section of the Western Society of Engineers in the auditorium of the latter organization. There were three hundred and seventy-five present.

"Significant New Developments in Ultra-High Frequencies" was the subject of a paper by G. C. Southworth, research engineer of the Bell Telephone Laboratories. This paper followed closely the one previously given in New York City which was summarized on page 261 of the March, 1938, issue of the PROCEEDINGS.

Two meetings were held in May, on the 6th and the 27th, at Fred Harvey's Union Station Restaurant, with J. E. Brown, chairman, presiding. There were one hundred and ninety-five at the meeting on the 6th and ninety-five at the other meeting.

At the May 6th meeting, W. L. Dunn, chief engineer of the Belmont Radio Corporation, presented a paper on "Variable-Inductance Tuning Applied to Automobile-Radio Remote Control." It covered the development of radio-frequency systems in which tuning is accomplished entirely by inductance variation caused by displacement of a powdered-iron core along the axis of a coil. The force required to move the core is much less than for a condenser gang. The system is well adapted to mechanical push-button control. Fixed capacitance in this system becomes a corollary of fixed inductance in a capacitively tuned system and fairly high-inductance cables may be employed between tuned circuits in the controlled head and tubes in the radio chassis. The tuning range is fixed by the ratio of maximum-to-minimum inductance. Oscillator design is similar to that with capacitive tuning; the same number of variables and degrees of freedom being possible and the same number of tracking "crossover points" are obtained. An analysis was presented of the noise characteristics of such design. The commercial applications of the system to automobile-radio equipment were described. The paper was discussed by Messrs. Hucke, Jarvis, Johnson, and Schorr.

The May 28 meeting was addressed by W. J. Polydoroff, consultant, who presented "Present and Future Developments of Television in Europe." It covered observations made by the speaker during recent travels abroad. The British television systems and standards were described. The Scophony system of projecting a picture by modulating a light source through the optical effects of pressure waves generated by a piezoelectric crystal in a transparent liquid was outlined. A projection tube employing two cathode rays was discussed. An account

was given of a French proposal for a reflecting screen to be used in a lighted room and on which the picture is formed by a mosaic of tiny shutters which control the black-and-white area of each picture element. The paper was discussed by Messrs. Bean, Brown, Johnson, and Kohler.

Cincinnati

The November 16 meeting of the Cincinnati Section was held at the University of Cincinnati with G. L. Platts, chairman, presiding. "The Fine Structure of Television Images" which appeared in the May, 1938, Proceedings was presented by A. V. Loughren of the Hazeltine Service Corporation.

The April 12 meeting was held jointly with the Dayton Engineers Club in their auditorium and with the Cincinnati and Columbus Sections of the American Institute of Electrical Engineers.

About two hundred attended the dinner at which W. H. Harrison, president of the American Institute of Electrical Engineers spoke.

A paper on "America on the Wire" was presented by C. H. Hoover of the Ohio Bell Telephone Company. In it he presented a history of the development of the telephone. Several sound films were shown and in addition, trips were made to Miller's Ford Station and the museum at Wright Field.

The annual meeting and dinner of the Cincinnati Section was held on May 24 at the Clovernook Country Club and was attended by thirty-nine members and guests. G. F. Platts, presided.

A. F. Knoblaugh, of the Baldwin Piano Company, presented a paper on "The Physics of Sound." Dr. Knoblaugh presented first a brief history of the development of the musical scale and musical instruments, particularly the piano. The harpsichord and clavichord, predecessors of the piano were described. It was pointed out that the equitempered musical scale made possible the playing of music written in different keys on the piano or organ.

The difference in tone qualities of various wind instruments was explained by reference to closed and open organ pipes. The closed pipe does not encourage the production of even-order harmonics whereas the open pipe accentuates them. This was followed by an outline of various principles of operation of electrical musical instruments and the manner in which they simulate the tones of the mechanical devices.

Connecticut Valley

The Connecticut Valley section met on April 28 in the Hotel Charles, Springfield, Mass. D. E. Noble, chairman, presided and there were forty present.

M. P. Wilder, engineer with the National Union Radio Corporation, presented a paper on "Circuit Simplification in Picture Generators." The subject was introduced with a discussion of the necessity for a picture generator for use in schools, demonstrations, development laboratories, or for supplying a signal for testing television receivers. A tube for providing this signal was described, and is called a monotron. It resembles a 3-inch cathode-ray oscilloscope except that the end, normally used for viewing, contains an aluminum plate on which the test picture is printed. The plate is held at seventy volts negative in relation to an aquidag coating on the inside of the tube. When the electron beam strikes the plate, secondary emission to the aquidag coating takes place leaving the plate momentarily positive. When the beam strikes the printed portion of the plate no emission occurs and thus a modulating voltage becomes available for transmission. A single tube generates a saw-tooth horizontal scanning sweep and the blanking and synchronizing voltages.

The use of plate-circuit filters was eliminated by employing a gaseous regulator tube. Pictures between three-hundred and six-hundred lines, sixty frames per second, sequentially scanned were available from this generator, and by simple modifications interlaced scanning could be obtained. The generator was demonstrated. The signal was picked up on an antenna on the other side of the room and fed to a British receiver which showed black-and-white pictures approximately eight inches square reflected from a front-silvered mirror mounted in the cabinet.

There were also described some projected television pictures produced by equipment developed in Germany and giving pictures about six feet square. The television camera used was of the image-dissector type and employed a 7-stage secondary-emission electron multiplier. On various stages of the multiplier the blanking impulses, carrier, and synchronizing impulses were inserted and there was obtained from the tube a complete signal ready for transmission at radio frequency. A general discussion followed.

Detroit

The Detroit Section met in the assembly room of WWJ, on April 22. F. S. Kaserman, vice chairman, presided and there were one hundred and twenty-five present.

G. C. Southworth, research engineer of the Bell Telephone Laboratories presented his paper on "Electromagnetic Waves in Free Space, in Metal Tubes, and in Dielectric Wires" which is summarized in the March, 1938, issue of the Proceedings.

E. H. I. Lee, chairman, presided at a meeting of the Detroit Section

on May 20. It was held in the Detroit News conference room and attended by eighty. V. J. Andrew, consulting radio engineer, presented a paper on "Directional Broadcast Antennas." It covered descriptions of several installations of directional antennas, and a step-by-step analysis of their design and operation was given. The theoretical and practical behavior of such systems were outlined. The paper was closed with a discussion of shunt-fed vertical radiators and the importance of ground systems.

An informal discussion led by the chairman was on present radio regulations and proposed changes in them.

Emporium

The following three meetings of the Emporium Section were held in the American Legion rooms and presided over by A. W. Keen, chairman.

The March 31 meeting was attended by forty and a paper on "Notes on Field-Intensity Measurements in the Broadcast Band and Their Interpretations" was presented by C. M. Jansky, Jr., of Jansky and Bailey. Taking three stations in a single city, the effect of frequency on coverage was shown; the lower frequencies are most effective in the daytime but less important at night. Antenna efficiency plays an important part in the coverage area and its pattern. Soil conductivity offers another important factor. The better soil conductivity of the Midwestern and South Central states is only offset by the lower density of population when coverage is compared with eastern stations. The fourth and least effective factor is transmitter power and, in general, is only about half as effective in increasing coverage as the other factors considered. It was pointed out that the relocation of a transmitter in an eastern city which permitted improved antenna design resulted in a one-hundred per cent increase in coverage for the same power. Coverage may be divided into zones which vary widely with night and day reception. For advertising purposes, the classifications of zones vary widely between rural and urban areas. The paper was discussed by Messrs. Abbott, Carter, and Jones.

R. F. Field, engineer for the General Radio Company, presented his paper on "The Schering Bridge" at the April 21 meeting of the section. This paper is summarized in the report of the Buffalo-Niagara Section meeting in this issue.

The May 26 meeting was attended by fifty-one and C. F. Miller, radio engineer of the Hygrade Sylvania Corporation, presented a paper on "A High-Frequency Amplifier Tube of New Design." The new

tube, designated type 1231, is a pentode designed for amplification at video frequencies. It combines high mutual conductance, low interelectrode capacitance, low self-inductance, stability of operation and physical design, and small space requirements. The electrical characteristics of the tube were presented and some of its mechanical and electrical features which are not readily apparent were discussed. Both internal and external shielding of the elements is employed to reduce capacitance. The leads are used as mount supports and base pins and a metal base shielding provides a positive locking of the base in the socket. One of the pins is used for the grid lead thus eliminating the top cap connection and the trouble encountered with flexible leads to it. The paper was discussed by Messrs. Atkins, Baldwin, McNally, and Mueller.

Indianapolis

The three following meetings of the Indianapolis Section were held in the Indianapolis Athletic Club. The first two were presided over by V. C. McNabb, chairman and the last by C. F. Wolcott, vice chairman.

The January 28 meeting was attended by twenty-two. David Seibert, loud-speaker engineer of the RCA Manufacturing Company, presented a paper on "Loud Speakers for Theaters and Public-Address Applications." The structure of various types of loud speakers was first discussed. Mathematical methods of obtaining the response-frequency and directivity characteristics were then outlined. It was shown that certain types of horns and baffles are superior to others and that the efficiency could be increased greatly by suitable coupling of the diaphram to the load. Cellular horns were described and their advantage in distributing sound more uniformly throughout a theater emphasized. It was pointed out that undesirable reverberation may be eliminated by the proper arrangement of sound projectors. After a general discussion, several types of speaker units and horns were examined.

The March 11 meeting was attended by twenty-six. V. J. Andrew, consulting engineer, presented his paper on "The Application of Directional Broadcast Antennas" which is described in the report of the May 20 meeting of the Detroit Section.

The April 8 meeting, attended by thirty-seven, was the annual meeting. As a result of the ballot, C. F. Wolcott, of Noblitt-Sparks Industries was elected Chairman; I. M. Slater of P. R. Mallory and Company, was named Vice Chairman; and B. V. K. French, also of P. R. Mallory and Company was elected Secretary.

"Mixer Considerations" was the subject of a paper by C. R. Hammond, application engineer of the Ken-Rad Tube and Lamp

Company. Current problems met with various types of frequency-converter tubes for use in superheterodyne circuits were discussed. Characteristics and limitations of various tubes were considered at length. Other problems met in the design of frequency converters were also discussed.

Los Angeles

On April 19, R. O. Brooke, chairman, presided at a meeting of the Los Angeles Section which was held in the RKO motion-picture studio at Hollywood and attended by sixty-five.

Three papers were presented and the first, by Bert Miller, sound engineer for Warner Brothers Studio, was on "Acoustics as Applied to Motion Pictures." In it Dr. Miller explained the problems confronting the sound engineer in regard to equalizing for motion-picture reproduction. The problem is complicated by the requirement of louder signals in theaters than are needed for radio reception, limitation of the frequency range of reproduction, and the amplitude and phase distortion which must be overcome. Peculiarities of the construction of the human ear and their effect on this problem were described.

"Radio as Used in Motion-Picture Production" was the subject of the second paper given by George Grignon, research engineer for Paramount Pictures, Inc. It covered chiefly the use of radio as a means of communication between various sections of a motion-picture company while on location. The equipment was described and its use illustrated by specific discussions of its application in the filming of several nautical pictures.

The third speaker, J. K. Hilliard, sound engineer for Metro-Goldwyn-Mayer Picture Corporation, discussed "Noise Reduction Methods in Recording." The methods described include matting, pre-equalization and postequalization which result in complementary recording, and push-pull methods.

The meeting was closed with three demonstrations. The first of these, which was of equipment for recording on film, was given by Max Schoemaker of the RCA Manufacturing Company. A recording truck containing a complete portable recording channel was displayed through the courtesy of RKO and Thomas Sharp. Equipment for recording and reproducing acetate records was demonstrated by M. E. Collins of the RCA Manufacturing Company.

Montreal

The April and May meetings of the Montreal Section were presided over by A. M. Patience, chairman, and held in the Engineering Institute of Canada auditorium.

The April 20 meeting was attended by forty and a paper on "Radio

Interference—Location, Suppression, and Control" was presented by H. O. Merriman, engineer in charge of interference for the Department of Transport. He pointed out that the investigation of interference began in Canada about 1924 when the National Research Council studied methods of locating sources of interference. He then described the latest equipment installed in cars for the location of interference sources. Methods of suppression were discussed and included such interfering devices as street-car compressor motors, trolley arcs and surges, power-line disturbances, and therapeutic machines. Wide enforcement powers are given the Department of Transport in the Canadian Broadcast Act but definite regulations have not yet been announced in order that more experience may indicate what demands would be reasonable. It has been found that peak values of interfering voltages are more significant than average values and this has resulted in the use of peak-measuring vacuum-tube voltmeters in this work. Discussion was presented of the problem of fixing responsibility in cases where the listener employs an inferior type of equipment or inadequate pickup and complains of interference. The discussion was participated in by Messrs. Hammond, Kelsey, Mitchell, Nixon, Oxley, Sillitoe, Smith, and Thompson.

The annual meeting was held on May 11 and in the election of officers, Sidney Sillitoe of the Northern Electric Company was named Chairman; A. B. Oxley, of RCA Victor Company, Vice Chairman; and R. E. Hammond of the Northern Electric Company was elected Secretary.

"Some Recent Developments in Frequency Modulation at 40 Megacycles" was the subject of a paper by E. H. Armstrong, professor of electrical engineering at Columbia University. He presented first the historical background from the original observation of frequency modulation in connection with arc transmitters in 1911 to recent developments. He described methods of obtaining frequency modulation and showed how the input signal is varied directly as the amplitude and inversely as the frequency in order to produce the desired phase shift. In his system, the phase shift was multiplied about 3000 times. For a modulation signal of 30 cycles the side-band frequency is about 100 kilocycles. At the high modulating frequencies, the side-band frequency is less. Distortion may be kept low at all except the very lowest frequencies and the response may be made essentially flat from 15 cycles to 20 kilocycles. Receivers for this system were described. By means of sound-film recordings, improvements in the signal-to-noise ratio of a frequency-modulated system as contrasted with an amplitude-modulated system were demonstrated. Pictures were shown of the new transmitter and tower which are being constructed at the

present time near New York City. The paper was discussed by Messrs. Fisher, Moore, Oxley, Rushbrook, Sillitoe, and Smith.

New Orleans

On April 14 a meeting of the New Orleans Section was held in the Association of Commerce Building. George Peirce, chairman, presided and there were ninety-one present.

"Marine Radiotelephony" was the subject of a paper by C. N. Anderson of the Bell Telephone Laboratories. In it he discussed types of equipment, installation, and methods used in ship-to-shore radiotelephone service. Several sound motion pictures were shown at the conclusion of the paper.

Philadelphia

The Philadelphia Section met on April 6 at the University of Pennsylvania. The meeting was presided over by A. F. Murray, chairman, and there were three hundred and twenty members and guests present.

V. K. Zworykin, director of electronic research of the RCA Manufacturing Company, presented a paper on "The RCA Television Project." Dr. Zworykin outlined the development of television from 1878 when Carey invented a selenium-cell system using a separate wire to transmit the current generated in each picture element. He then discussed the history and theory of the iconoscope and kinescope as used for television.

Several new experimental iconoscopes make use of the electron-multiplying system which gives promise of increasing the sensitivity by ten or more times. The use of secondary-emission image multiplication was discussed. The optical image to be transmitted is projected on to a semitransparent photoelectric cathode and the resulting electron image is in turn focused on to a mosaic which may be single or double sided. In the latter case the image is projected on one side and scanned from the other. The picture signal in the form of secondary electrons emitted by the scanning beam can also actuate a secondary-emission multiplier.

Kinescopes have been improved by the development of new fluorescent materials and increased control of the cathode-ray beam by magnetic focusing. A description was given of the television studio located in the RCA Building in New York City, the connecting radio link operating at 177 megacycles, and the transmitters and antennas being located on top of the Empire State Building. Two trucks, comprising mobile television pickup stations for outside work were described. One truck carries the iconoscope camera and a parabolic microphone, while the other contains a transmitter and antenna for radiation at 177 megacycles.

A. F. Murray, chairman, presided at the May 5 meeting which was held at the Engineers Club and attended by two hundred and thirty.

The first paper "A New High-Mutual-Conductance, High-Frequency Amplifier Tube" was presented by A. P. Kauzmann of the research and engineering department of the RCA Manufacturing Company. This is a new high-grid-plate-transconductance pentode applicable to high-, intermediate-, and video-frequency amplification. The high transconductance is obtained through the use of closely spaced tube dimensions. It was pointed out that the effect of the close spacing of cathode and control grid in the RCA-1851 provides an increase in g_m which more than offsets the effect of the increase in input capacitance. Electrical and mechanical characteristics of the tube were shown. Because of its high transconductance with correspondingly low plate current, the input noise is lower than in conventional tubes. Input loading and capacitance variations caused by plate-current changes can be minimized by using an unby-passed cathode resistor of about 40 ohms. Gains of between two and four have been obtained in a tuned high-frequency stage operating at 50 megacycles. At 11 megacycles gains of twelve to twenty-five have been observed with a band-pass of 4.5 megacycles. The paper was discussed by Messrs. Hervey, Somers and others.

J. S. Starrett, of the research and engineering department of RCA Manufacturing Company, presented a paper on "Nonradio Application of Radio Tubes." Various general uses of vacuum tubes were first described and the processes of making phototubes outlined. Among special applications were a photoelectric equipment to control the roasting of coffee through color detection, a device using a curtain of light in conjunction with phototubes for safety purposes, an automatic automobile-engine valve-stem-tempering control, and a gaseous beam tube for measuring magnetic intensity and polarity. Among various tubes which were demonstrated were special glassware filled with inert gas which glowed when brought within the field of a high-frequency oscillator. Devices for measuring the speed of a passing automobile by photoelectric means and a burglar-alarm system operating over the house-wiring circuits by means of carrier currents were described. The paper was discussed by Messrs. Bingley, Murray, West and others.

Pittsburgh

On April 19, the Pittsburgh Section met in the Carnegie Institute of Technology, R. T. Gabler, chairman, presiding. There were fifty-five present. A paper on "The Schering Bridge" by R. F. Field, described in the report of the Buffalo-Niagara Section, was presented.

San Francisco

On April 6 the San Francisco Section held a meeting at Manning's Coffee Cafe with C. J. Penther, vice chairman, presiding. There were fifteen present.

A paper on "Beam Power Tubes" by O. H. Schade, published in the February, 1938, Proceedings, was reviewed by Robert Sink. Alex de Bretteville, Jr., led the discussion on "The Effects of Space Charge in the Grid-Anode Region of Vacuum Tubes" by Salzberg and Haeff, which appeared in the January, 1938, issue of the *RCA Review.*

"Some Recent Developments in Demountable-Tube Design" was presented by Charles Litton of the Litton Engineering Laboratories at the April 27 meeting of the San Francisco Section which was held in the Heintz and Kaufman plant. Noel Eldred, chairman, presided and there were fifty-nine present.

The speaker presented first a brief historical outline of the development of demountable power vacuum tubes. He then discussed specific tubes developed in his laboratories and built for power ratings above 10 kilowatts. Each tube is provided with a 2-stage mechanical displacement pump followed by a 2-stage oil-vapor pump with a charcoal trap, all assembled as a compact unit. A 100-kilowatt tube with its associated pump equipment can be manufactured for practically the same cost as the regular 100-kilowatt tube. Filament replacements in the demountable tube can be made at a cost of about 3 per cent of the original tube and pump cost. It was pointed out that another economy factor was the use of high-efficiency filaments because of the possibility of frequent replacements. The paper was closed with a demonstration of a demountable tube and its pump equipment.

The "Student Paper" meeting of the San Francisco Section was held on May 18 in the Pacific Telephone and Telegraph Company auditorium. Noel Eldred, chairman, presided and there were thirty-two present.

F. C. Cahill of Stanford University presented a paper on "The Application of Negative Feedback to Harmonic-Analyzer Design." The second paper was by R. W. Hill of the University of California and covered "An Impedance Bridge for Measurements at a Frequency of 200 Megacycles." The judges awarded the prize of a year's Associate membership in the Institute to Mr. Cahill.

President Pratt attended the meeting and gave a brief review of the recent conference in Cairo, Egypt. He commented also upon some of the interesting meetings at the World Radio Congress held in Sydney, Australia.

Seattle

"The Present Status of Television" was the subject of a paper by Austin V. Eastman, professor of electrical engineering at the University of Washington which was presented on April 28 at a meeting of the Seattle Section held at the University. A. R. Taylor, chairman, presided and there were fifty-two present.

Professor Eastman reviewed first the elementary theory of television. The necessity of wide transmission bands to obtain acceptable definition was pointed out and its effect on requiring high carrier frequencies stated. These high carrier frequencies result in limited coverage and expensive operation except in metropolitan areas. The greater population density in European countries makes economic television coverage more feasible than in this country. Descriptions were given of the various television systems and their methods of operation. The paper was discussed by Messrs. Fisher, Hurlbut, Kiebert, Libby, Walker, Wallace and others.

Toronto

The following three meetings of the Toronto Section were held in the University of Toronto and presided over by W. H. Kohl, chairman.

The April 11 meeting, which was attended by seventy-four, was addressed by M. A. Acheson of the Hygrade Sylvania Corporation on the subject "Short Cuts in Vacuum-Tube Design." He indicated first the need in vacuum-tube-design engineering for comparatively simple formulas to designate the changes in the various characteristics of tubes as a result of alterations in the shaping and size of the various elements. He then presented a number of such formulas which assisted materially in arriving at suitable designs with only a small amount of computation. The paper was discussed by Messrs. Bayly, Kohl, and Parker.

H. M. Smith, design and construction engineer of the Canadian Broadcasting Corportation, and W. H. Doherty, engineer in the radio development department of Bell Telephone Laboratories, presented a paper on "CBL" at the April 26 meeting. There were eighty-eight present.

Mr. Smith presented first the problems encountered in choosing a suitable site for CBL. Various factors such as ground conductivity, population, cost of land, power supply, communication lines, and general accessibility were considered. The relation between antenna height and field intensity was next considered and it was decided that a height of 0.57 wavelength was desirable. The use of a grounded vertical radiator gives greater protection from lightning and simplifies

the problem of illuminating the tower. A ground system comprising 240 wires is employed. A coaxial transmission line runs from the station equipment to the tower and is connected approximately 0.15 wavelength from the bottom of the tower. It was pointed out that in the design of the coaxial line, the mathematical factor e gives the optimum ratio of the diameters of the inner and outer conductors. He then briefly discussed the station equipment and pointed out that the Doherty system made possible a saving in power of about $5000 per year.

Mr. Doherty then described the transmitter which employs grid modulation in one of the low-power stages. Negative regeneration is employed by taking some of the power delivered by the final amplifier, rectifying it, and feeding it back to the grid of the first audio-frequency amplifier, thus producing a substantial improvement in hum level and over-all distortion. The 50-kilowatt class B amplifier was then described. It employs the Doherty circuit in which one tube is adjusted for maximum efficiency at zero modulation. Its plate is connected to the plate of the second tube through a π network and the input impedance of this network varies inversely with the output load, the phase shift through the network being ninety degrees. To compensate for this shift, the input to the grid of the first tube is also shifted ninety degrees and this permits the second tube to contribute power on modulation peaks only. The operation was demonstrated by the use of a small transmitter and cathode-ray oscillograph. The papers were discussed by Messrs. Bayly, Irwin, Parker and others.

Toronto Section

The annual meeting of the Toronto Section was held on May 9 at the University of Toronto. W. H. Kohl, chairman, presided and the attendance totaled sixty.

A paper on "Circuit Design and its Relation to Tube Performance" was presented by L. C. Hollands of the RCA Manufacturing Company, Radiotron Division. It consisted of a series of problems which have arisen in connection with circuit design. The circuits discussed included oscillator circuits for converter tubes, load circuits for power output tubes, inverse-feed-back circuits, battery-type audio-frequency amplifiers with special references to grid-blocking problems under certain switching conditions, microphonics caused by variations in heater-cathode capacitance in certain types of oscillators, hum in fixed-bias audio-frequency amplifiers, and zero-bias operation of audio-frequency tubes.

Messrs. Choat, Dawson, Hackbusch, Hepburn, and Parker participated in the discussion.

In the election of officers, R. C. Poulter, of the *Radio Trade Builder*, was named Chairman; G. J. Irwin, of Philco Products, Ltd., Vice Chairman; N. M. Potter, of Canadian National Carbon Company, Ltd., was selected as Secretary-Treasurer; and H. S. Dawson, of Rogers Radio Tubes, Ltd., Recording Secretary.

Washington

At the April 18 meeting of the Washington Section, held in the Potomac Electric Power Company auditorium, and presided over by E. H. Rietzke, chairman, Robert Field presented his paper on "The Schering Bridge." This is summarized in the report of the Buffalo-Niagara meeting elsewhere in this issue. The meeting was attended by ninety-five.

Chairman Rietzke presided at the May 9 meeting of the Washington Section held in the Potomac Electric Power Company auditorium and attended by seventy-seven.

R. M. Wise, chief radio engineer of the Hygrade Sylvania Corporation, presented a paper on "A High-Frequency Amplifier Tube of New Design." This is the type 1231 tube which is described in the report of the May 26 meeting of the Emporium Section. The paper was discussed by Messrs. Bailey, Rietzke, Wheeler and others.

Personal Mention

J. M. Borst has joined the radio engineering department of the Aerovox Corporation of Brooklyn, N. Y., having previously been with *Radio News*.

P. H. Dorte has left Gaumont-British Picture Corporation to join the British Broadcasting Corporation as production manager of television outside broadcasts.

E. K. Kelley, previously with the RCA Manufacturing Company, has joined the engineering department of the Ken-Rad Corporation at Owensboro, Ky.

Barton Kreuzer has been transferred from New York City to Hollywood, California, by the RCA Manufacturing Company.

B. S. Longfellow, formerly with Hygrade Sylvania Corporation, has joined the Federal Communications Commission staff as inspector, in New York City.

E. W. Sanders has left the RCA Manufacturing Company to become radio inspector with the Federal Communications Commission at Fort McHenry, Baltimore, Md.

Proceedings of the Institute of Radio Engineers
Volume 26, Number 7 — *July, 1938*

TECHNICAL PAPERS

TELEDYNAMIC CONTROL BY SELECTIVE IONIZATION WITH APPLICATION TO RADIO RECEIVERS*

By
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Summary—*This paper concerns a new and improved method of remote control which may be applied to electrical devices of any sort. Special attention is given to the exposition of the basic principles of the system as applied to the remote control of radio receivers.*

The new control arrangement utilizes the alternating-current power line as a medium for transmission of control signals from a remote unit to the controlled device, and is therefore free of the limitations imposed by control cables.

The system utilizes a small oscillator, located at the remote point, operating at a frequency between 200 and 400 kilocycles. Use of the 60-cycle line voltage as plate supply causes the radio-frequency output voltage of the remote unit to be modulated in intensity. Control signals are transmitted through the alternating-current power line to the controlled point, where the phase of the 60-cycle line voltage is compared with the phase of the envelope of the radio-frequency control signals in any one of several phase detectors which may be employed for selective purposes. Relays in the detector output circuits are energized and may be made to close power circuits to any desired electrical device.

Remote control of the start of operation of an electric appliance may be had without the expenditure of stand-by power in receptive circuits at the controlled point. This is accomplished with a cold-cathode gas tube in a circuit which is responsive to a given remote-oscillator carrier frequency, whose modulating envelope is compared in phase with the 60-cycle line voltage. A gas discharge is initiated, and this closes a relay whose contacts connect the terminals of the electrical appliance to the line.

Various combinations of carrier frequency and envelope phase may be impressed on the power line at the remote point to cause desired functions to take place at the controlled device. Two different carrier frequencies, with two different phase relations of the modulation envelope, may be arranged in combination to result in ten different functions at the controlled point. If three frequencies and two envelope phase arrangements are used, the number of possible combinations, and consequently, different functions which may transpire, is increased to twenty-six.

An experimental remotely controlled receiver is described, and it is shown that the use of two carrier frequencies and two envelope phase arrangements at the remote

* Decimal classification: R570. Original manuscript received by the Institute, December 11, 1937. Presented before Rochester Fall Meeting, November 10, 1937; New York Meeting, December 1, 1937; Pittsburgh Section, February 15, 1938.

point, makes it possible to turn the receiver on or off, to select any one of six stations, and to increase and decrease the volume level.

Introduction

MANY electrical devices of the type used commonly in the home and in industry are particularly adaptable to remote control. The application of remote control to home broadcast receivers appears to be particularly pertinent at this time, because it is the next logical step in the sequence of events related to the development of tuning aids, of which push-button control or so-called "electric tuning" is the latest feature.

Many systems of teledynamic control have been applied to radio receivers in the past, but none has achieved its intended aim, principally because of three basic limitations. One of these is the necessity for employing intricate mechanical tuning arrangements in the receiver, in order to attain the required precision in tuning. In addition to this, the use of an interconnecting cable is considered objectionable because it limits the range of operation to within the length of the cable and because cables are not acceptable on the floors of homes. The consumption of power by control circuits during nonoperating periods of the receiver is the third reason for the failure of remote-control systems in the past. This stand-by power is generally required if remote-control of the "on-off" feature is desired.

The ideal remote-control system, which could be applied to radio receivers or any other device ordinarily controlled by electrical means, would have the following characteristics. In the first place, it would require no additional wiring circuits; the control signals would be transmitted from the remote point to the controlled device over circuits already in use for some other purpose, or, possibly, by radiation through space, with no interconnecting wires. The control power required would be small, and, above all, there would be no continuous power consumption during nonoperative periods of the controlled device. Power would be supplied from the control unit only during the performance of the desired control functions. The control should be decisive, and should not be affected by spurious interference from power circuits or radio signals. In addition it should be possible to operate several similar remote-control arrangements independently and without interaction, and also without causing interference in radio-receiver circuits.

Recognition of the need for an improved system of teledynamic control led this laboratory to investigate the possibilities of developing a system which would fulfill the requirements of the ideal remote-control

method as previously outlined. It was intended that the new system should eliminate many of the limiting factors present in preceding arrangements, and should, if possible, combine some of their better features.

The general plan of attack was made apparent, and the undertaking was given considerable impetus, when it was found that the alternating-current power line could be used as a medium for the transmission of currents of low radio frequency, that is, less than 500 kilocycles. The possibility of using the power wires as an interconnecting link for the transmission of control signals between the remote and controlled points was recognized immediately, and an investigation of the power-line characteristics was undertaken.

I. Nature of the Problem

The general problem involving transmission of carrier currents on alternating-current supply lines has several aspects. First, the properties of the alternating-current line as a transmission medium must be determined, and the attenuation and line impedance at several frequencies must be known. Similarly, the power level required for the transmission of control signals is important, since the size and cost of apparatus are directly concerned.

The attenuation is probably the most important property of the alternating-current line when it is used as a carrier for radio-frequency currents. If the attenuation is high, the power output of the control-signal source used as the remote-control unit must necessarily be quite large; this requires considerable equipment, in the form of power oscillators and amplifiers at the remote point.

The power dissipated in the control circuits at the controlled device is another factor which determines the required power capabilities of the remote-signal source, since, even with low line attenuation, the control power applied to the line at the remote point is a function of the energy consumed at the controlled point. This latter quantity will be determined generally by the type of circuit element employed to utilize the control energy. If power amplifiers were operating continuously to afford amplification of the control signals at the controlled point there would be no necessity for high power levels at the remote unit. This would permit using the amplified control signals directly to perform the desired functions, or they could be employed to cause relays to close certain power circuits. In this latter case the control energy from the line could initiate some secondary action, which might take the form of plate current passing through a relay in a power de-

tector. The energy used to actuate the desired device could then be drawn directly from the detector power supply.

In case stand-by power is considered objectionable, one might use a type of control device which could be actuated directly by the control-signal energy. For instance, a copper-oxide rectifier might be used, but its operating power consumption is generally too large.

The choice of the proper frequencies for control purposes, and the alternating-current line impedance at these frequencies, are both quite important. The frequencies employed must lie in a certain rather limited band, because of possible interference with broadcast signals in the radio receiver. The line impedance should not be too low, if proper impedance matching at the terminating points is to be accomplished readily. Similarly, it should not be too high, for high-impedance lines are affected in their transmission capabilities by variable power loads.

II. Transmission and Power Considerations

Line impedance and attenuation were measured over the frequency range from 200 to 400 kilocycles. This is the most useful range because its upper end is nearly the highest frequency which can be transmitted efficiently over the power line. Also these frequencies lie outside any bands used for broadcast purposes, and also outside the intermediate-frequency bands of radio receivers; this is an essential requirement, if interference in the receiver from the control signals is to be avoided.

The line attenuation to these frequencies proved to be not very large. The maximum voltage attenuation on power lines in this laboratory was less than 10 to 1, as measured between the most remotely separated power outlets.

The average results of many line-impedance measurements made in typical buildings wherein a teledynamic control system of this type might be installed are as follows:

Average line impedance at 200 kilocycles $= 25 + j40$ ohms
Average line impedance at 300 kilocycles $= 40 + j60$ ohms

Loading the line with soldering irons, lamps, and similar low-impedance appliances has little effect on the line attenuation, because of the low impedance of the line itself.

The maximum line attenuation at frequencies of 200 and 300 kilocycles is about 20 decibels. This means that a signal source at the remote point with a power output of one watt would supply a resultant power of 0.01 watt at the controlled end of the alternating-current line, even under the most unfavorable attenuation conditions.

The possibility of using only a watt or so of radio-frequency energy at the remote point is interesting, for, with this power rating, the remote unit may be made quite small physically. Since this unit will generally contain a radio-frequency oscillator as a source of control energy, a low-power requirement makes the use of a small receiving-type tube quite feasible.

One hundredth of a watt available at the far end of the line may seem to be insufficient for control purposes, but it can be shown that it may be utilized to control, in many ways, nonpower-consuming devices. Assume that the frequency of the control signals is 300 kilocycles, and that the alternating-current line is properly terminated at the controlled point in an impedance of 72 ohms (the resultant of $40+j60$ ohms, the average line impedance at 300 kilocycles). Then the radio-frequency voltage available at the far end of the line is about 0.8 root-mean-square volt. The use of a resonant circuit, tuned to 300 kilocycles, will result in an increase in amplitude of this voltage, due to resonant rise, to 30 or 40 volts root-mean-square, and at this level it may be applied to various circuits for control purposes.

The rise in voltage is not accompanied by any gain in power, consequently the increased voltage cannot be used to operate a power-consuming device, such as a copper-oxide rectifier, because the tuned circuit across which the voltage appears is inherently a high-impedance arrangement. The radio-frequency power drawn from the line should be used only to initiate some control action, whose energizing power is supplied from the 60-cycle power line.

A power amplifier might be so arranged that its output current could be controlled by the 30 or 40 radio-frequency volts available due to resonant rise in the tuned circuit connected across the line. A relay, energized by the amplifier output current, could then be used to close any desired power circuit. This arrangement requires that the amplifier be in continuous operation, receptive at all times to control signals emanating from the remote point. The stand-by power consumed is somewhat objectionable, and requires that frequent replacement of control tubes be made due to their continuous operation, even during quiescent periods of the receiver. Accordingly, some means is needed for starting the controlled devices from the remote point without requiring the use of intervening amplifiers in continuous operation.

III. Basic Remote-Control Circuit

Several tube and circuit arrangements were investigated to find a way to make this "cold-starting" feature possible. The requirements of no stand-by power eliminated immediately the use of a tube (such as a

diode) with a heater continuously operating, and pointed the way to the selection of a cold-cathode type of gas tube as the vital circuit element for use in this work.

The basic arrangement finally selected involves a gas tube and associated circuit elements which are so disposed that a small radio-frequency voltage transmitted over the power line to the pickup circuits and increased in amplitude by resonant rise in a tuned circuit, may be employed to add to the potential of one of the gas tube's elements and thereby to initiate a gas discharge; this, in turn, causes a relay to close, connecting the controlled device to the 60-cycle power line as a source of energy.

Fig. 1 shows the constructional details of a gas tube which adequately fulfills the requirements. The tube consists of three elements,

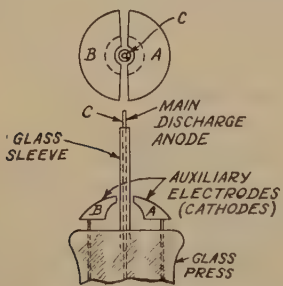

Fig. 1—Constructional details of gas tube.

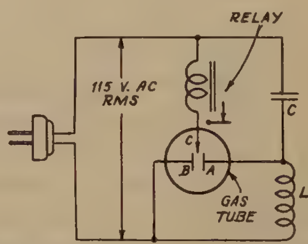

Fig. 2—Basic gas-tube circuit.

enclosed in an envelope containing gas at low pressure. The two similar elements, which are called auxiliary electrodes, are symmetrically disposed with respect to the third element. The two auxiliary electrodes act as cathodes, since the third element can act only as an anode to receive electrons from the other two. This is due to its construction and orientation with respect to the cathodes, which gives the tube a rectifying characteristic.

In a gas tube having the construction shown in Fig. 1, the gap between A and B becomes ionized if the potential difference across it exceeds 70 volts. This is true regardless of the polarity of the voltage difference. Once started, the discharge is maintained as long as the potential difference exceeds 60 volts. The potential required to initiate a discharge between the anode C, and either of the cathodes A or B, in the absence of ionization in the gap A–B, is of the order of 200 volts. The discharge in gaps C–A or C–B, once started, is maintained if C is 60 volts or more positive with respect to either of the two cathodes. If gap C–A or gap C–B is ionized due to the application of the required

breakdown voltage, a low-impedance current path exists in the main gap only if C is positive with respect to either of the cathodes. This action is due to the rectifying characteristic mentioned above.

The main gap will become ionized at 70 volts on C if first the gap between A and B is broken down. Thus, the potential on C may be made any value less than 200 volts and no discharge in the main gaps C–A or C–B will occur until gap A–B is broken down. When this happens, a large current can be made to flow in the anode circuit if the anode is more than 70 volts positive with respect to one of the cathodes when the auxiliary gap A–B becomes ionized.

One way to utilize the controlling effect of the radio-frequency voltage transmitted from the remote unit is shown in Fig. 2. Here the gas tube is the same as that described above. The main gap C–B is sup-

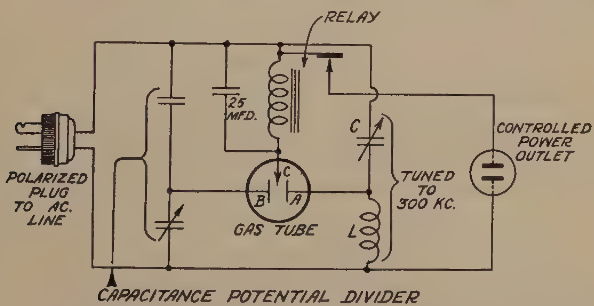

Fig. 3—Basic gas-tube circuit with means for improving sensitivity.

plied with 60-cycle voltage from the power line, but, since the peak voltage is only $\sqrt{2}$ times the line voltage, or about 180 volts, there is no discharge, because of the 200-volt threshold breakdown potential in the main gap.

If, however, a volt or two of radio-frequency energy from the remote unit is present across the power line in Fig. 2, it is increased in amplitude by resonant rise in the series circuit LC, which is tuned to the carrier frequency of the remote oscillator. The resultant radio-frequency voltage which appears across L, and consequently between A and B, may be over 70 volts peak; this will cause gap A–B to become ionized. The potential required to ionize the main gap is thereby decreased to 70 volts, and current flows through the relay winding during the portions of the alternating-current cycle when C is more than 60 volts positive with respect to A.

The amount of radio-frequency voltage required to initiate the discharge in gap A–B may be reduced, and the sensitivity of the device increased, by supplying element B with a 60-cycle potential as shown

in Fig. 3. The main anode current then passes between C and A through L, instead of from C to B, because of the large 60-cycle reactance of the capacitance connected between B and ground. The capacitance voltage divider may be adjusted to apply 50 peak 60-cycle volts between B and A, thereby reducing the radio-frequency voltage required to break down gap A–B to about 20 volts. The sensitivity could be further increased by applying a still larger "polarizing" 60-cycle potential to B, but this would make the device susceptible to line voltage variations.

The use of a capacitive voltage divider, in place of a resistive potentiometer, for setting the polarizing potential of element B, permits the operation of the circuit with zero stand-by power.

The discharge in the main gap is not continuous, with operation on alternating current as shown in Fig. 3. This would be true even if the auxiliary gap were continuously ionized. The reason for the intermittent nature of the main discharge lies in the sinusoidal variation of the voltage on element C, for, as soon as the instantaneous value of this voltage becomes less than 60 volts positive with respect to A, the main discharge ceases, and is reinitiated only when the potential of C rises again during the next cycle to 70 volts. It is seen that the main discharge is not maintained even intermittently if the radio-frequency voltage is removed from coil L, since the voltage across gap A–B is then insufficient to reinitiate the auxiliary discharge (which is also intermittent in nature) during the next cycle.

If several functions are to be performed, several gas-tube circuits, each receptive to a different carrier frequency, could be used at the controlled point. There is, however, a unique expedient at hand which may readily be used to decrease the number of carrier frequencies required. This method of control makes use of a radio-frequency voltage, modulated in intensity by the 60-cycle line voltage at the remote unit, and a comparison of the phase of the line voltage at the gas-tube circuit with the phase of the modulation envelope of the radio-frequency voltage transmitted over the power line.

The source of energy for this type of remote control comprises a conventional feed-back oscillator, whose plate-supply voltage is derived from the 60-cycle line, instead of from a direct-current source. Application of alternating voltage as plate supply permits the tube to oscillate only in the portion of the alternatingcurrent cycle during which the oscillator tube's plate is positive with respect to its cathode. Consequently, the oscillator output voltage, which is fed into the line at the remote point, is modulated in intensity, and consists of a series of pulses as shown in Fig. 4. The intervals during which os-

cillation takes place may be shifted by 180 degrees at 60 cycles if the phase of the line voltage is reversed before being applied to the oscillator circuit. This could be done by reversing the remote-unit power-supply plug. The radio-frequency impulses then take the position shown dotted in Fig. 4.

The radio-frequency voltage, modulated in intensity, is picked up at the line terminals in Fig. 3 and appears across coil L as previously explained. The main gap C–A will not be ionized in this case, even with sufficient radio-frequency voltage available across coil L, unless the phase of the modulation envelope is properly related to the phase of the alternating-current line voltage on element C. The reason for this "selective ionization" is that the radio-frequency voltage adds to the

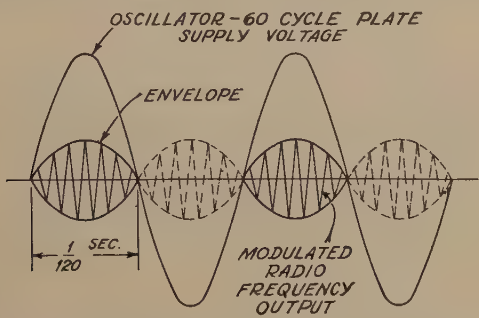

Fig. 4—Modulated output of remote oscillator.

60-cycle voltage across gap A–B only during a half cycle at 60 cycles; hence, the gap A–B is ionized only during a portion of this interval. If the auxiliary discharge occurs during the portion of the alternating-current cycle when element C is negative with respect to A, the low-impedance discharge in the main gap does not take place and the relay is not energized. Under these conditions a reversal of the phase of the radio-frequency envelope would cause ionization between A and B, when C was positive with respect to A, and the main-anode discharge current would be sufficient to actuate the relay.

It can be seen that the gas tube must have a rectification characteristic if phase control of this type is to be possible, otherwise the relay will be energized on either phase of the radio-frequency envelope.

IV. Use of Selective Ionization for Control Purposes

Initiation of the gas discharge in a tube connected in a circuit similar to that of Fig. 3 may be controlled by two variables; one is the carrier frequency of the remote oscillator, and the other is the phase of the radio-frequency oscillator output-voltage envelope. Several simi-

lar gas-tube circuits may be set up at the controlled device and may be controlled independently from a remote oscillator, in which both the carrier frequency of the radio-frequency voltage and the phase of the modulation envelope are variable.

Assume that two oscillators in a remote unit may be made to operate at either 200 or 300 kilocycles, and that the 60-cycle modulating phase of either oscillator may be reversed. The controlled unit may then contain four gas-tube circuits similar to that shown in Fig. 3. The resonant circuits associated with two of these would be tuned to 300 kilocycles, with the 60-cycle potentials on the tube elements so phased that the individual tubes are ignited by oppositely phased radio-frequency envelopes. The second pair of gas tubes would be arranged in the same manner for actuation by phase difference in the

TABLE I

Oscillator No. 1		Oscillator No. 2		Gas-Tube and Relay Number
Frequency	Envelope Phase	Frequency	Envelope Phase	
300	A			1
		300	B	2
200	A			3
		200	B	4
300	A	300	B	1 and 2
200	A	200	B	3 and 4
300	A	200	B	1 and 4
200	A	300	B	2 and 3
300	A	200	A	1 and 3
300	B	200	B	2 and 4

modulation envelopes, but their resonant circuits would be tuned to 200 kilocycles.

Let each gas tube have a relay in its anode circuit, and let the relays be numbered respectively 1, 2, 3, and 4. It is then possible to show, as is done in Table I, that ten different combinations are available by using the two oscillator frequencies (200 and 300 kilocycles) and the two opposite phases of the radio-frequency envelope, *A* and *B*.

The four relays may, therefore, be closed singly or in combinations of two at a time, depending upon the conditions existing at the remote unit. The relay contacts can be used to connect any electrical device to its source of power or to close other control circuits.

If three carrier frequencies were used, with both phases of the modulation envelope, the number of possible combinations of frequency and phase would be increased from ten to twenty-six. This would require six gas tubes at the controlled point, each with a multiple contact relay in its anode circuit.

It is to be noted that the relays remain closed only as long as the radio-frequency voltage is supplied to the line. For simple "on-off"

switching an arrangement may be made to eliminate the necessity for maintaining the radio-frequency voltage during operating periods of the controlled device. This can be done by using two gas tubes, whose resonant circuits are tuned to the same frequency, but which are ignited by opposite phases of the modulating radio-frequency envelope. One tube is used to close a small relay, which, in turn, closes a locking relay, whose contacts make the power circuit for the controlled device. The other gas tube is arranged to be actuated by the opposite phase of the radio-frequency envelope, and its relay is made to open the locking relay, thereby breaking the connection between the controlled device and the power line.

V. Other Types of Phase and Frequency Detectors

A gas tube is not the only type of electronic device which may be used as a phase detector in this method of remote control. High-vacuum receiving-type tubes may readily be arranged in conjunction with other circuit elements to provide the phase detection required for separation of the various control signals. However, since vacuum tubes require power in their heater and plate circuits, it is necessary to use a gas tube (as a power-supply switch) in conjunction with the vacuum tube if "cold-starting" properties are to be realized.

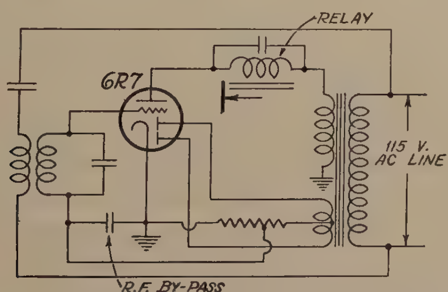

Fig. 5—Phase detector utilizing high-vacuum tube.

One example of the application of high-vacuum tubes for phase detection is shown in the circuit of Fig. 5, which includes a tube of the 6R7 type; this tube contains a triode and two small diodes. Plate-supply potential is obtained from the alternating-current power line, either directly or through a step-up transformer. The radio-frequency control signals are separated in frequency and applied to the grid of the tube by means of the tuned circuit, which is connected to ground in series with a suitable bias source.

The function of the bias is to maintain the tube at cutoff in the absence of control signals. This may be done by applying to the grid a

60-cycle voltage which opposes the alternating-current plate voltage in phase, and is slightly greater than the plate voltage divided by the amplification factor of the tube. No plate current flows under these conditions, but the grid current may be excessive during positive excursions of the grid alternating voltage, if the alternating-current source has low impedance. To reduce this grid current, one may insert a high resistance in series with the grid to cause rectification during positive alternating-current grid pulses, which action will have no effect on the detecting properties of the circuit.

Another way to maintain the tube at plate-current cutoff is to use a full-wave rectified, unfiltered 60-cycle voltage developed in the double-diode portion of the tube, as shown in Fig. 5. The voltage is applied as negative grid bias. The ratio of the amplitudes of the plate and grid voltages is somewhat less than the μ of the triode section. There is no grid current under these conditions, hence no series grid resistor is necessary.

The detector circuit operates in the following manner: Plate current flows through the relay winding when the phase of the radio-frequency envelope is so arranged that the peak instantaneous grid voltage occurs coincidentally with the peak positive alternating plate voltage. If the radio-frequency envelope is shifted 180 degrees in phase at the remote-control point, no current flows in the phase-detector plate circuit because the plate is negative with respect to cathode (due to the 60-cycle plate voltage) when the pulse of radio-frequency voltage is applied to the grid.

The sensitivity of the device may be increased by rectifying the modulated radio-frequency voltage present across the tuned circuit and extracting the low-frequency modulation components. The output of the diode rectifier used for this purpose is applied to the grid in series with the bias source. The increased sensitivity is due to the fact that the average value of the grid voltage is increased due to rectification in the diode.

The circuit of Fig. 5 has the necessary qualifications for selection of the proper control signals, since it has both phase and frequency selectivity. Extension of this arrangement to include relay actuation on either phase of the modulation envelope may be made by using an additional triode, whose alternating plate voltage is in phase opposition to that on the tube described. More direct, however, is the use of a double triode of the 6N7 type, as shown in the circuit of Fig. 6. Here the alternating plate voltages are in phase opposition, and the relay which is actuated depends upon the phase relation between the radio-frequency envelope and the alternating-current plate potential. Ob-

viously both relays will be energized if both phases of the modulation envelope are fed into the line at the remote point.

A double triode connected for phase detection as described above may be used in conjunction with one carrier frequency; two double triodes may be substituted for the four gas-tube circuits described in the first part of the paper, and operation with two frequencies and both phases of the radio-frequency envelope results in ten combinations

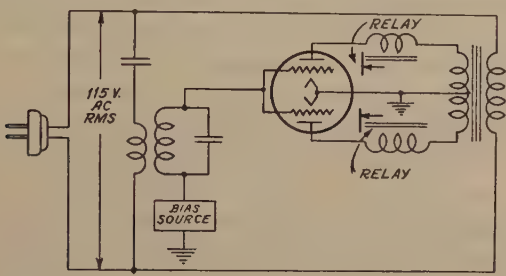

Fig. 6—Double-triode high-vacuum phase detector.

of phase and frequency, which will actuate four multiple-contact relays as described above. This, of course, would eliminate the "cold-starting" feature.

Small thyratrons or gas triodes may be substituted for high-vacuum tubes in the phase detector if it is desired to operate some power-consuming device directly from the anode current of the tube. In this way a high-current solenoid or relay may be controlled by causing the gas discharge to be initiated by the properly phased radio-frequency envelope.

VI. Remote Control of Radio Receivers

One of the principal uses of this method of remote control lies in its application to radio receivers. Recognition of the importance of this aspect of the general problem led to the development of an experimental remote-control receiver, in which the control circuits previously described were utilized. This development served as a field test to determine the capabilities of the method under actual operating conditions.

The experimental receiver is a standard model, which was selected principally because it was equipped with automatic frequency control. This feature is a necessary adjunct to any receiver tuned by mechanical means, since no tuning mechanism available at the present time is sufficiently precise in its action to permit tuning with the required degree of accuracy.

The nature of the control system is such that, by means of control circuits heretofore explained, the same functions are performed electrically from a remote point as have been accomplished in previous systems which utilized a motor and contactor arrangement, wherein the control is effected by push buttons, located either at the receiver or removed therefrom by a cable connecting the receiver with the remote point.

The relays in the control circuits, acting singly or in combinations of two at a time, cause the motor power circuits to be closed, due to the fact that the relay contacts are arranged to parallel the push-button switches which normally control the motor operations. The alternating-current motor is used to vary the tuning or volume level, while the receiver is turned on with the gas-tube control circuit described above.

The motor unit, which is similar in type to most arrangements used for remote-control tuning purposes, contains a 23-volt capacitor-type alternating-current motor, which is provided with a tapped reactor and condenser for changing the phase of the armature current so that rotation in either direction is possible. The armature is connected, through gear trains, either to the shaft of the tuning condenser bank or to the volume-control potentiometer. When the motor operates at 18 volts the armature rotates in a position to engage gears connected to the volume control potentiometer. 23-volt operation causes the armature to move axially to engage gears connected to the tuning-condenser shaft.

The direction of rotation and the points at which the motor stops are controlled by a series of cams and contactors. The cams, mounted on a shaft coupled to the shaft of the tuning condenser, rotate as the tuning is changed, and actuate the contactors, which are, in effect, single-pole double-throw switches. These make or break the circuit between the motor and its supply, depending upon the position of the contactor rider on the periphery of the cam. There are six cams of this type on the motor used in the model receiver, and each is provided with means for preselection of any desired station.

VII. Details of the Control System

There are ten different functions which must be performed in the receiver from the remote point. The receiver must be turned on and off, any one of six stations tuned in, and the volume level increased or decreased. These functions are all controlled by the characteristics of the energy fed into the line at the remote point. The ten functions are accomplished by using two carrier frequencies, 200 and 300 kilocycles, and two opposite phase arrangements of the radio-frequency envelope.

The circuit of the remote unit is shown in Fig. 7. The two triodes (contained in a single 6N7) are so arranged, in conjunction with push-button switches and a center-tapped power transformer, that either tube may be made to oscillate at 200 or 300 kilocycles, with either or both of two possible phases of the modulating envelope. Each of the nine push-button switches, when depressed, cause different combinations of frequency and envelope phase to be fed into the power line. Only nine switches are used to perform ten functions; this is possible because the "on" switch is used for volume-increase action once the receiver at the far end of the line is turned on.

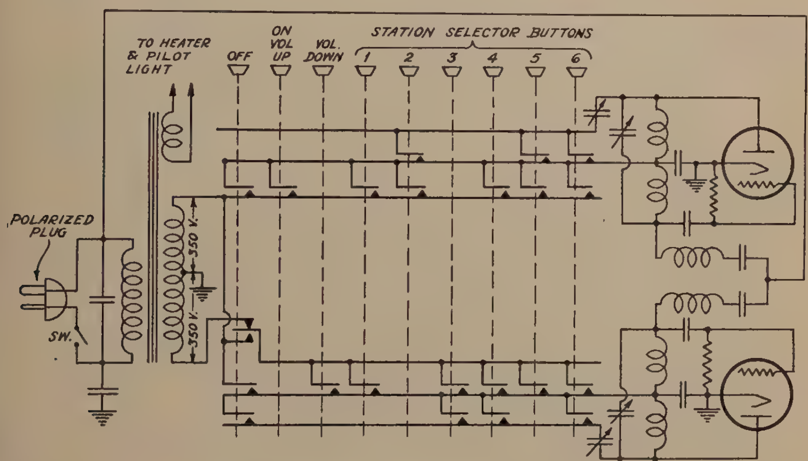

Fig. 7—Circuit of remote-control unit.

Fig. 8 shows the details of the receiver-control unit. This includes a gas-tube circuit, a single-sided phase detector with means for developing a rectified bias voltage (6R7), and a double-phase detector of the 6N7 type.

The gas-tube circuit is used to turn the receiver on and is actuated by 300 kilocycles modulated in, say, phase A (envelope phase). When relay No. 1 closes, due to a discharge in the gas tube, a locking relay is also made to close, thereby connecting the receiver power supply and control unit to the alternating-current power line. When the locking relay closes, relay No. 1 is left free for other purposes, thereby obviating the necessity for maintaining the discharge in the gas tube while the receiver is operating.

The 6R7 acts as a phase detector, with a resonant circuit tuned to 300 kilocycles connected between its grid and cathode. The 60-cycle voltage on its plate is so phased that if 300 kilocycles modulated in

phase B (opposite to that required to initiate the gas discharge) is applied to the line, plate current flows and relay No. 4 is energized. The diode section is used to develop the full-wave rectified 60-cycle voltage used as bias for the two vacuum-tube detectors.

The 6N7 serves as two phase detectors in one envelope. Here the frequency of the tuned grid circuit is 200 kilocycles and the alternating-current plate potentials are supplied from a center-tapped power

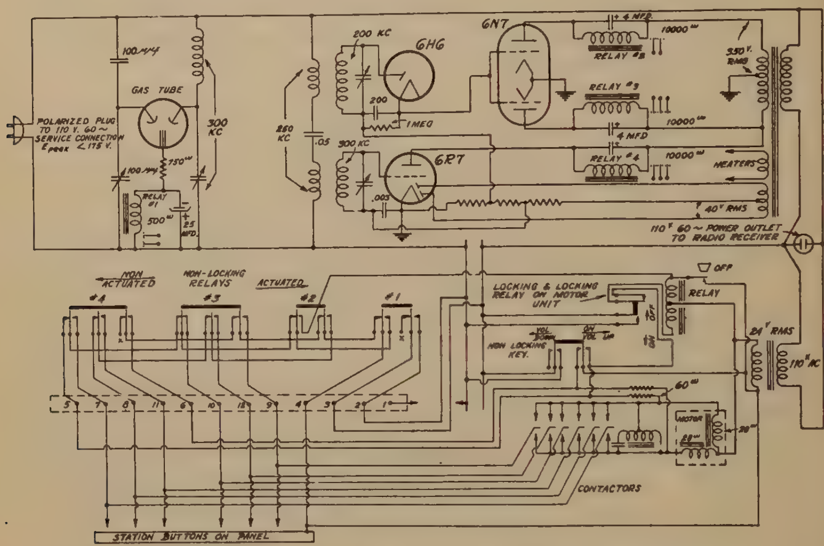

Fig. 8—Circuit of receiver-control unit.

transformer. The top tube passes current and energizes relay No. 2 when 200 kilocycles modulated in phase A is applied to the line at the remote point. Reversal of the modulation envelope to phase B causes plate current to flow in the lower tube and relay No. 3 is closed.

The 6H6 used in conjunction with the 200-kilocycle tuned circuit is not an inherent part of this control system, for it is used merely to extract the low-frequency component of the envelope of the 200-kilocycle voltage by ordinary diode detection. The diode output voltage is applied to the grids of the 6N7 connected in parallel. This results in a larger change in direct plate current because of the increase in average value of the rectified envelope as compared with that of the modulated 200-kilocycle voltage. A tube with higher mutual conductance than the 6N7 would obviate the need for using the 6H6.

There is no stand-by power consumed in the control unit during nonoperative periods of the receiver. The power consumed while the

receiver is operating and no control functions are being performed is only that required in the heater circuits of the 6N7 and 6R7, and in the diode-rectifier section of the 6R7. Both vacuum tubes are biased to cutoff in the absence of control signals from the line, hence no plate-circuit power is consumed.

The oscillator unit dissipates power only in its heater circuits when no control functions are being performed; this is about five watts.

The relays are shown in Fig. 9 in their nonactuated positions. Their contacts are arranged to parallel the push-button switches on the front of the receiver.

The receiver may also be controlled manually by means of the motor, which is actuated by the six push buttons mounted on the front panel. These control the tuning operation. Volume control is effected by means of a toggle switch which causes the motor to operate at reduced voltage to engage the volume-control gears.

Three of the four multiple-contact relays used in the control unit have 10,000 ohms resistance and are actuated by about 5 milliamperes current. The other relay (used in the gas-tube circuit) has about 500 ohms resistance and operates on about 30 milliamperes.

VIII. Interference Considerations

The possibility of interaction between two installations of this type connected to adjacent power circuits must be considered. Generally, the two remote-control arrangements will be isolated for radio-frequency currents due to the watt-hour meters and associated terminal apparatus and the attenuation between the two power circuits will be sufficiently high to preclude interaction of the two installations.

There may, however, be instances in which the natural attenuation is not sufficiently high to permit independent operation. In these cases means must be taken to isolate mutually the respective power circuits for the radio-frequency signal currents. This may be done by inserting radio-frequency chokes at the meter boards of each power circuit. If this is not feasible, different sets of control frequencies may be used in the two remote-control arrangements; for instance 200 and 300 kilocycles for one installation and 250 and 350 kilocycles for the other.

Radiation of energy from the power lines at the carrier frequencies used in this remote-control system has been found, by measurement, to be negligibly low. Consideration of the factors which affect the radiation substantiates this experimental data.

In the first place, the carrier-frequency power developed in the remote oscillator unit is not more than one watt. Second, the power wires act very inefficiently as radiators. This is particularly so in

modern electrical installations in which the alternating-current lines are placed in metal conduit, which greatly reduces the radiating possibilities.

Harmonic radiation also is low because the circuits used to couple the remote oscillator to the line are designed to prevent the transmission of harmonics from the oscillator to the line.

The only other source of radiation is the tank circuit of the remote oscillator. This has been found to be small, and it can be eliminated entirely in practice by shielding the oscillator unit. In other words, the problem, practically speaking, involves the same considerations as do receivers of the superheterodyne type.

Proceedings of the Institute of Radio Engineers
Volume 26, Number 7 *July, 1938*

THE RELATION BETWEEN RADIO-TRANSMISSION PATH AND MAGNETIC-STORM EFFECTS*

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Summary—*This paper presents the results of a quantitative study of the relationship between the proximity of great-circle transmission paths to the magnetic pole and of signal stability during terrestrial magnetic disturbances. Reception from Europe as observed at Riverhead, Long Island, and San Juan, Puerto Rico, is compared during normal and disturbed periods. The great-circle path from Europe to San Juan is about 1000 miles farther from the north magnetic pole than a similar path to Riverhead.*

A brief description of the duplicate equipment and antenna systems employed at the two locations is included.

It has been known for some time that the more southerly transmission paths are apparently less susceptible to effects accompanying magnetic disturbances. The results of observations covering a period of years confirm the anticipated relationship between signal stability and proximity of the transmission path to the magnetic pole. The average advantage of San Juan over Riverhead for European signals during disturbed periods is found to be approximately 8 decibels.

Evidence of higher ionospheric ionization over the more southerly path is noted in a number of cases by a comparison of nighttime cutoff effects and allied phenomena.

Introduction

NUMEROUS investigators have recognized the existence of a relationship between radio field intensity and the proximity of the transmission path to the earth's magnetic pole.[1,2,3,4,5,6] It is the purpose of this paper to report some of the results of observa-

* Decimal classification: R113.5. Original manuscript received by the Institute, February 4, 1938. Presented before Silver Anniversary Convention, New York, N. Y., May 12, 1937.

[1] H. E. Hallborg, "Terrestrial magnetism and its relation to world-wide short-wave communications," Proc. I.R.E., vol. 24, pp. 455–471; March, (1936).

[2] R. M. Morris and W. A. R. Brown, "Transoceanic reception of high-frequency telephone signals," Proc. I.R.E., vol. 21, pp. 63–80; January, (1933).

[3] C. R. Burrows and E. J. Howard, "Short-wave transmission to South America," Proc. I.R.E., vol. 21, pp. 102–113; January, (1933).

[4] C. R. Burrows, "The propagation of short radio waves over the North Atlantic," Proc. I.R.E., vol. 19, pp. 1634–1659; September, (1931).

[5] C. N. Anderson, "Notes on radio transmission," Proc. I.R.E., vol. 19, pp. 1150–1165; July, (1931).

[6] H. H. Beverage and H. O. Peterson, "Diversity receiving system of R.C.A. Communications, Inc., for radiotelegraphy," Proc. I.R.E., vol. 19, pp. 531–561; April, (1931).

tions initiated in 1934 to study the magnitude of this effect quantitatively.

Utilizing transmissions from several European stations, it was proposed to study the increase in immunity from magnetic disturbance obtainable by removing the point of reception from Riverhead, Long Island, to a point as far south and east as practicable. An examination of the great-circle map for London shown in Fig. 1 emphasizes that little or no advantage may be secured by relocating the receiving point within the southern United States, since the great-circle path for points in southern Florida differs but little from that of Riverhead, and the static level is, of course, much more adverse in southern latitudes. By removing the point of observation to the West Indies, however, it is possible to secure a transmission path about 1000 miles more remote from the north magnetic pole. The island location also contributes to a lower noise level. Rio Piedras (near San Juan), Puerto Rico, was therefore selected as a site approximating the optimum conditions obtainable among the possessions of the United States. A slight further advantage could be secured by the use of the Virgin Islands as a receiving location. However, the difference is small, and the site chosen was particularly convenient, since it permitted a collaborate program with the University of Puerto Rico, which has been most co-operative in providing facilities for the carrying out of the Puerto Rican observations.

General Considerations

Zones of Maximum Magnetic Disturbance

Several investigators have pointed out that radio signals are, in general, less affected by terrestrial magnetic activity when the direction of propagation approaches a north-and-south line. For example, signals traveling east and west, or traversing the earth's field perpendicularly, are much more subject to fluctuations in intensity during magnetic storms than are signals traveling north and south, or parallel to the earth's field. Signals traversing the higher northern latitudes, especially in the vicinity of the magnetic pole, are particularly erratic in their behavior.

Since the earth's magnetic field is greatly concentrated at the north and the south magnetic poles, it is reasonable to assume that any changes in the total field would be intensified in the magnetic polar regions. This has been determined actually to be so. Magnetic disturbances are found to increase in intensity with north latitude. Hallborg,[1] through correlation studies of radio-circuit performance and horizontal-intensity fluctuations, has set up criteria for deter-

mining the commercially dead and disturbed zones for any communication center.

In Figs. 1 and 2, respectively, are shown the zones of maximum magnetic disturbance for London and Moscow. It should be understood that there is actually no line of demarcation between the disturbed and undisturbed zones; the transition from one to the other is gradual. The size of the sector defining the limits of the disturbed zone

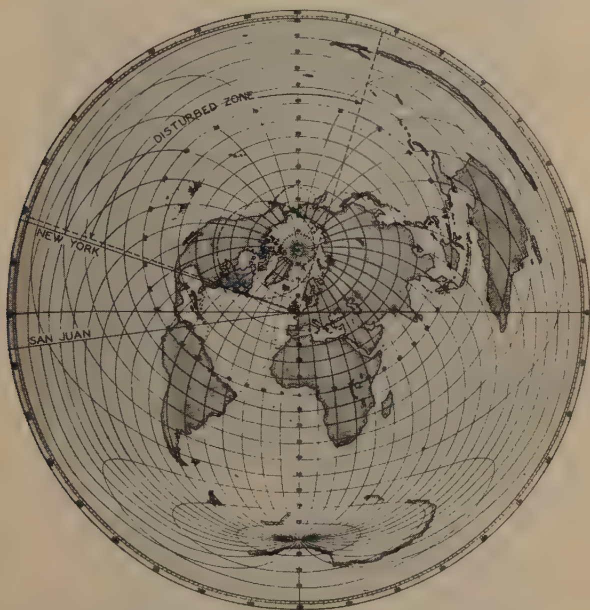


Fig. 1—Azimuthal map of the world around London, showing magnetically disturbed zone according to Hallborg's criterion, and the marked advantage of the San Juan path over the New York path.

is arbitrarily determined from the average of horizontal magnetic intensity ranges for moderate magnetic storms. It depends upon the relative limits of circuit disturbance selected and hence upon the type of communication service. Furthermore, the size of the zone, being dependent on magnetic variability, will increase and decrease with the eleven-year cycle of sunspot activity. For the radius of the circle about the magnetic pole, which fixes the boundaries of the disturbed sector, we have chosen a value equal to 30 degrees of latitude. This value will of course hold for only a specific period of the sunspot cycle. However, it will be the same for all transmitter locations and so can be used in making valid comparisons of transmission paths.

An inspection of Figs. 1 and 2 will show that the divergence in

transmission paths from Moscow to Riverhead and to San Juan is not as great as in the case of the paths from London. In view of this, a

Fig. 2—Azimuthal map of the world around Moscow. New York and San Juan paths and their relation to the magnetically disturbed zone are shown.

smaller advantage for San Juan over Riverhead would be expected for signals from Moscow. This is found to be the case.

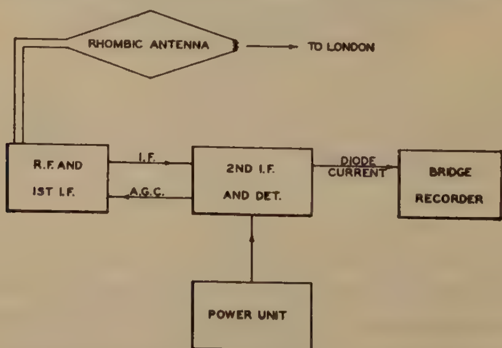


Fig. 3—Block schematic of signal-intensity recording systems used at Riverhead and San Juan.

Description of Equipment

For this investigation, four identical receivers adapted for automatic recording were constructed at Riverhead, Long Island. These receivers are of the triple-detection type and provide an exceptionally high and measurable intermediate-frequency amplification. To sim-

plify their construction, existing commercial units, modified to suit special requirements, were used as components.

A schematic diagram of the system employed is shown in Fig. 3. The physical appearance of the recording equipment is shown in Fig. 4. Leeds and Northrup Wheatstone-bridge recording galvanometers of the old type are utilized to record signal levels. These are especially suited to this kind of work; the used portion of the recorder

Fig. 4—The two recording assemblies employed at San Juan.

paper can hang through an opening in the bottom of the machine, making it easy to cut off each day's record. Also, the records from these recorders can be readily blueprinted, thereby permitting a complete file of the records obtained at both points of observation to be retained at each observing point.

After completion and preliminary testing, two of the receivers (completely assembled in the panel form shown in Fig. 4) were shipped to San Juan, Puerto Rico, and installed in a laboratory constructed at the University of Puerto Rico.

The site of the Puerto Rican measurements was a clear field about a half mile from the campus of the University of Puerto Rico and far removed from active roads and sources of man-made interference. A special building was constructed for this program. This building and its environs are shown in Fig. 5. The site and building facilities at Puerto Rico were made available by the University of Puerto Rico which cordially co-operated in the program. The Riverhead measurements were taken at the RCA Communications receiving site at Riverhead.

Fig. 5—Building constructed for Puerto Rican measurements. One of the supporting poles for the rhombic antenna can be seen at the left.

Rhombic antennas of identical design were constructed in Puerto Rico and at Riverhead. These antennas are 400 feet on the long diagonal and 214 feet on the short diagonal and were designed for maximum response at 14 megacycles. The antennas are directed toward London. Some of the 60-foot poles supporting the rhombic antenna used in Puerto Rico are visible in the background of Fig. 5.

Standard-signal generators are used to provide frequent calibrations of the sets in operation at both points. These were tested and compared before the equipment was shipped to Puerto Rico. The generator used at Puerto Rico is shown in Fig. 4.

The equipment used at both points of observation was made as far as possible identical, so as to facilitate a comparison of absolute

signal levels at the two points of observation as well as to study level changes during magnetically disturbed periods. In this paper, however, we shall be concerned primarily with a comparison of the magnitude of the changes in level over various paths during magnetic storms, and our conclusions are hence not dependent on the absolute calibration of the recordings.

Scope of Observations

The major series of observations studied cover transmissions received during the past two years from the British stations GLK (8005 kilocycles) and GLH (13,525 kilocycles), both located at Dorchester (London), England, and RKB (15,625 kilocycles), located at Moscow, Russia. The British signals were chosen because of their stability and consistent operation; the Russian signal was utilized, because its more northerly path affords particularly interesting transmission problems.

Additional series of shorter duration were taken on some other stations to gain a better idea as to the extent the results secured were applicable to other paths and frequencies. It will be noted that the results of these observations are in good accord with the conclusions derivable from the longer series.

Results of Observations

Interpretation of Data for Different Stations and Paths

In interpreting the available transmission data, involving over one thousand observation days distributed among about six stations, it is convenient to adopt parameters which facilitate the intercomparison of data taken on different stations and at different periods.

Magnetic observatories commonly adopt a 0, 1, and 2 basis for rating the magnetic character of a day, i.e., 0—undisturbed, 1—moderately disturbed, 2—severely disturbed (magnetic storm). In this paper we avoid confusing seasonal and longer-period phenomena with short-period magnetic effects by confining ourselves to a consideration of the reduction from monthly 0-day signal-level averages caused by disturbances indicated by days of magnetic character 1 and magnetic character 2. When a longer period of observations is available, it is planned to consider in another paper the seasonal and longer-period phenomena, which are already becoming evident. A discussion of these changes will necessitate a consideration of directional transmitting-antenna characteristics and the absolute calibrations of receiving equipment to render significant the conclusions derivable. The results derived in this paper are rendered independent of all of

these variable factors by the use of decibel increments from monthly averages, which permit ready and valid intercomparisons between data obtained on stations varying in output power and antenna direc-

TABLE I

Average Reductions in Signal Levels for Disturbed Days, Compared with Averages for All Undisturbed Days in the Series

Station	Character number of respective days in series			Average db reductions for character-1 days		Average db reductions for character-2 days		Average db advantage for San Juan	
	0	1	2	River-head	San Juan	River-head	San Juan	1 days	2 days
GLH 13,525 kc	473	205	13	5.1	−0.1	12.2	0.8	5.2	11.4
GLK 8005 kc	243	104	3	8.5	1.2	10.3	−3.9	7.3	14.2
RKB 15,625 kc	218	96	10	5.8	3.8	19.8	14.4	2.0	5.4
LCP 14,550 kc	33	26	2	6.7	−0.5	9.9	1.7	7.2	8.2
GNZ 4250 kc	15	4	0	8.0	−0.3			8.3	
DGG 13,180 kc	2	4	0	15.0	4.0			11.0	

TABLE II

Average Reductions in Signal Levels for Disturbed Days, Compared with Undisturbed-Day Monthly Means for Months in which Disturbed Days Occurred

Station	Character number of respective days in series			Average db reductions for character-1 days		Average db reductions for character-2 days		Average db advantage for San Juan	
	0	1	2	River-head	San Juan	River-head	San Juan	1 days	2 days
GLH 13,525 kc	473	205	13	4.6	0.3	11.2	2.3	4.3	8.9
GLK 8005 kc	243	104	3	7.2	1.0	8.0	1.3	6.2	6.7
RKB 15,625 kc	218	96	10	6.5	4.0	17.0	11.4	2.5	5.6
LCP 14,550 kc	33	26	2	4.5	−1.7	7.5	3.5	6.2	4.0
GNZ 4250 kc	15	4	0	8.0*	−0.3*			8.3*	
DGG 13,180 kc	2	4	0	15.0	4.0			11.0	

* The GNZ series embraced a period extending into two successive months. Only a portion of the total undisturbed days were in the same month as the disturbed days, and, as a result, computations made according to the uniform method employed in this table give values which are clearly not representative. Since the total number of observation days is less than a month (19 days), the "month" in this case has been regarded as covering the entire period.

tivity characteristics. Twenty-four-hour decibel averages, or decibel averages extending over as long a period as simultaneous observations permit, are utilized.

Comparative Susceptibilities for All Frequencies and Paths Observed

Tables I and II summarize the average decibel reductions in signal level for all observed stations for days of magnetic character 1 and days of magnetic character 2. In Table I the averages for all 1 and 2 days are compared with the grand averages for all 0 days in the respective periods of observation. In Table II the deviations for 1 and 2 days are compared against the 0-day monthly means for only the particular months in which the disturbed days occurred.

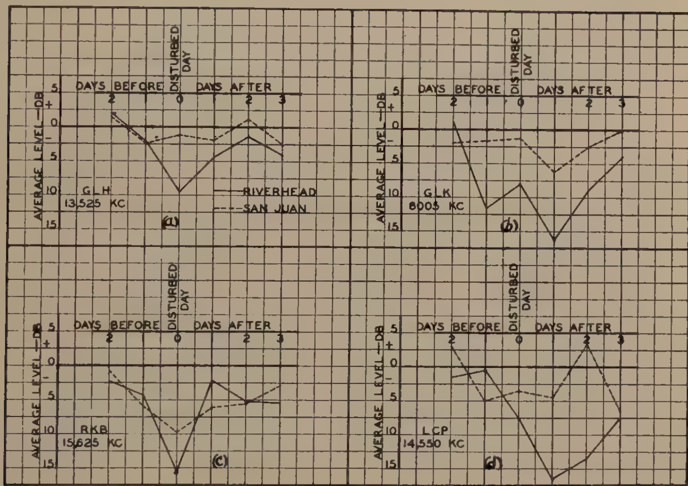


Fig. 6—Decrease in average daily signal level for severely disturbed periods for various stations observed at Riverhead and San Juan. Note relative immunity of the southern-path signals to these disturbances.

It is believed that the method of comparison used in Table II is to be preferred. By referring disturbed days to the undisturbed days of the same month in which they occur, the inclusion of varying seasonal effects is, to a great extent, avoided. In any series extending over a considerable length of time, the daily signal levels do not by any means remain constant. They are subject to seasonal and longer-period variations. Comparing conditions during an abnormal period with average conditions holding for a short time contiguous to the disturbance tends to present a truer representation of the actual state of affairs, since the results then are not influenced by figures prevailing when altogether different conditions may exist. There is some tendency to reduce slightly the magnitude of the recorded effect, because of possible recovery lags in signal levels extending into 0-day periods of magnetically active months. It is believed, however, that this is

of minor importance. A fair agreement between values obtained from Tables I and II attests the fact that the conclusions are not appreciably affected by such factors. It will be noted that the durations of the series are quite variable, but that the magnitude of the effects indicated for the stations reported does not show evidence of appreciable sampling errors for the shorter series, except perhaps in the extreme case of DGG.

Figs. 6(a) to 6(d) show comparisons of the form of the deviation in field intensities for the stations observed in days contiguous to days of magnetic character 2. It will be noted that in a number of cases,

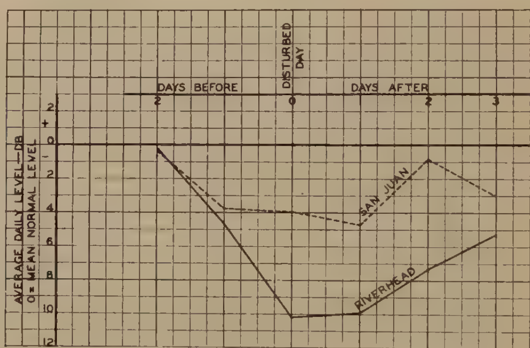


Fig. 7—Average decrease in average daily signal level for severely disturbed periods for all stations observed. Note relatively greater decrease in Riverhead levels.

the influence in the San Juan field is so slight as to render the form of the San Juan curves subject to sampling errors which are eliminated only by a grand average for all stations and all magnetic days of character 2. This is shown in Fig. 7. In Fig. 6 the 2-day averages are grouped where more than one occurs on a given station. In some of the series only one or two 2 days were available, so the desirability of an average curve such as shown in Fig. 7 for the elimination of undue sampling errors is evident.

Fig. 8 gives a summary of the effects of individual magnetic storms. Here are shown the reductions in average signal level at Riverhead and San Juan for all signals observed during severely disturbed periods. It has been found that the maximum depressing effect on radio transmission does not necessarily occur simultaneously with the maximum magnetic activity, but has a tendency to lag behind it. Occasionally, the full effect is not felt until the following day. When this situation takes place, the average signal intensities are sometimes increased,

rather than decreased, on the 2 day. In Fig. 8 the instances when maximum signal depression occurred on the day following the magnetic 2 day have been indicated.

Since we are here interested mainly in evaluating the radio effect of any magnetic storm, regardless of time of occurrence, the values used in the preparation of Fig. 8 are more suitable for drawing conclusions than are those of Table II. In this table (as in Table I) the

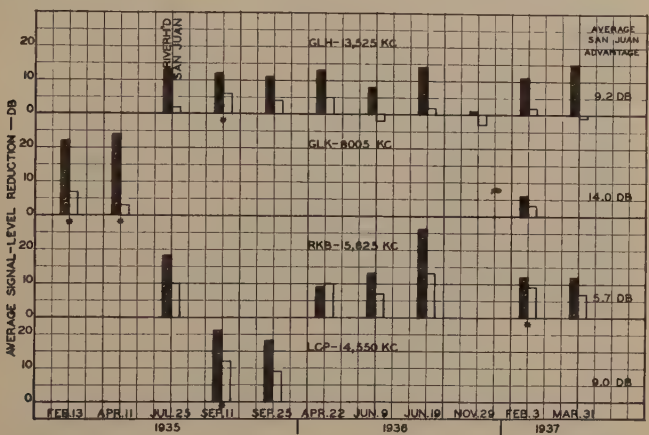


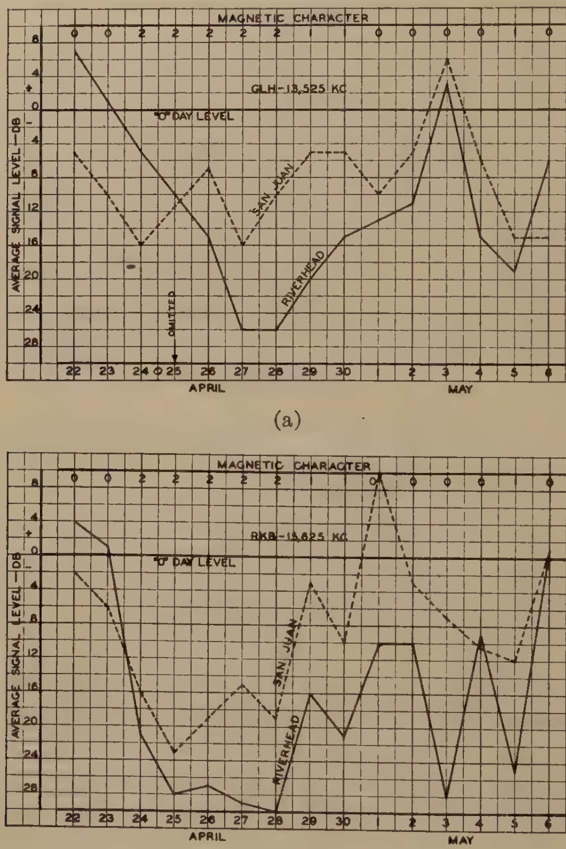
Fig. 8—Summary of individual magnetic storms, showing maximum daily-average signal-level reductions at Riverhead and San Juan for all signals observed.

* Indicates that maximum signal depression came on day following date shown, which is day of maximum magnetic activity.

magnetic effect is given primary consideration. The signal reductions for all disturbances being classified according to the magnetic character of the day on which they took place, the actual effect on radio transmission remains masked to some extent. Because of the lag in the arrival of maximum signal depression the maximum reductions caused by severe disturbances must at times necessarily be included in the 1-day averages. In view of this, and coupled with the fact that there are occasional actual increases in intensity on some 2 days, it will be evident that calculations involving a consideration of only the 2 days will give average intensity values which are higher than they really should be. The result of this is to lower the differential advantage of the southern path. Comparison of the 2-day averages in Table II and the average values shown in Fig. 8 will give an idea of the magnitude of such discrepancies. The averages of Fig. 8 have been assumed to give the correct values for the severe disturbances.

The average differential reduction for a magnetic storm for the

British stations, GLH and GLK, and for the Norwegian station, LCP, is found to be approximately 10.4 decibels; that for the Russian station, RKB, is about 5.7 decibels. As would be expected, the effect for LCP is of the same order of magnitude as that for the British stations; the divergence in transmission paths is approximately the same. For



(a)

(b)

Fig. 9—Reduction from April 0-day averages of signal level produced by the extreme disturbance of the week of April 23, 1937.

RKB, where the path divergence is considerably less, the differential advantage for the southern circuit is not nearly as pronounced.

The data included in Figs. 6 and 7 embrace the period from January, 1935, to March, 1937. During this period, no successive days of magnetic character 2 occurred, so those figures may be considered to represent the disturbance produced by the ordinary magnetic storm of moderate intensity.

Beginning April 24, 1937, there occurred a magnetic storm of extreme severity and duration. During this period several 2 days followed in succession.

Plots of the field-intensity reductions produced on this occasion are shown in Figs. 9(a) and 9(b). Some idea of the severity of the disturbance is shown in Fig. 10, where the relative reductions are shown for this period and for the previous two years. It will be noted that the larger reductions show a similar differential in favor of the more southerly path.

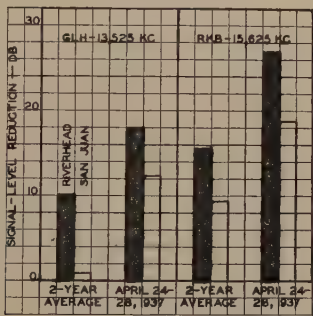


Fig. 10—Comparison of Riverhead and San Juan 2-day reduction in signal level for the period from January, 1935, to March, 1937, against reduction for week of April 23, 1937.

Relationship of Magnitude of Disturbance to Path and Frequency[7]

While it will be noted that the form and magnitude of the effects shown in Fig. 6 are on the whole surprisingly consistent, yet they display certain phenomena which at first sight appear contrary to what one would anticipate from *a priori* consideration. Thus, the reduction of the field indicated for GLK (8005 kilocycles) over the more northerly path is greater than that for GLH (13,525 kilocycles). It is believed, however, that this effect is probably due to a diversity in the period of observations including the two series. The GLK period of observation included winter periods during which GLK was employed for night communication. During the early winter months of 1935, this frequency was near the critical frequency at these hours over the

[7] The data for Figs. 9 and 10 were not reduced in time for inclusion in the original draft of the paper as presented at the I.R.E. convention on May 12, 1937. The discussion under this section does not include these additional data. They were considered as extreme examples and are perhaps better omitted from a general discussion of frequency and seasonal changes in the susceptibility of transmission to magnetic storm effects. Tables I and II have been modified to include the additional data.

paths under consideration. Inasmuch as the London—San Juan path extends farther to the south, the critical frequencies encountered over this path during the late evening periods in the winter of 1934–1935 were distinctly higher than for the London—Riverhead path. This resulted in a late evening advantage for San Juan in this period as compared to Riverhead (see Fig. 11). It has already been suggested[4] that the effect of a magnetic disturbance is to lower critical frequencies so that the advantage of San Juan as compared to Riverhead was particularly

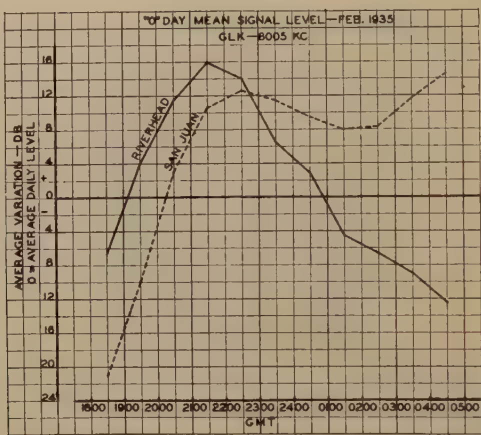


Fig. 11—Average variations from average signal level for GLK (8005 kilocycles) at Riverhead and San Juan for undisturbed days of February, 1935. Puerto Rican fields were much better sustained during late evenings, indicating cutoff phenomena on the northern path.

evident when average fields for a period including night hours was considered. This is further clarified by a comparison of the curves shown in Fig. 12 for the February, 1935, storm with the curve for the undisturbed period shown in Fig. 11. The observations on GLH and most of the other stations operating above 13 megacycles were initiated at a later period in 1935 and during 1936 when the critical frequencies were abnormally high, caused, no doubt, by changes in the eleven-year sunspot cycle. This resulted in the frequencies under measurement being further below the critical at both points of observation, despite the higher absolute frequencies. This situation is in accord with the fact that during much of the period of observations it was possible to operate these frequencies all night at Riverhead without fade-out. Under such conditions, it appears that the advantage of the more southerly path is of smaller magnitude.

Factors Influencing Susceptibility of Transmission Path to Magnetic Disturbance

The relationship between magnetic disturbances and the proximity of the great-circle transmission path to the magnetic pole has already been discussed by several observers, but a complete understanding of the mechanism of the connection appears to await more comprehensive data relating the ionization of the ionosphere with latitude, and particularly the effect of magnetic storms upon this relationship. From

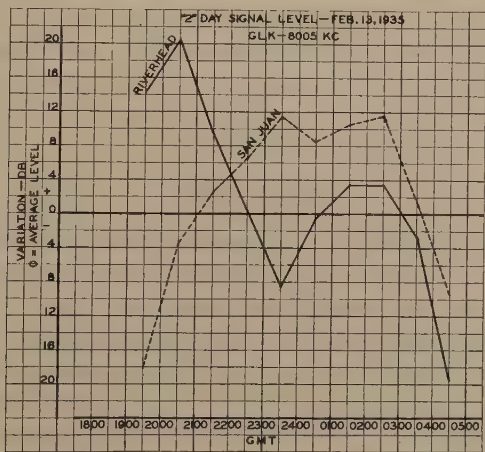


Fig. 12—Variations from average signal level for GLK (8005 kilocycles) at Riverhead and San Juan during severely disturbed day of February 13, 1935. Note increased early afternoon and evening Riverhead fields and late evening decrease for San Juan, due to lowering of cutoff frequency.

the limited data so far available relating to this subject, it appears that two major phenomena relevant to our problem are associated with the magnetic phenomena. Thus, the effect of a magnetic storm upon high-frequency radio transmission appears to result from increased absorption in the higher ionized layers, which is apparently most evident in close proximity to the poles. In addition to the increased absorption, which may possibly be caused by diffusion, a decrease in critical frequencies also accompanies these disturbances. The first of these phenomena may react to accentuate this effect, as in the case of the 1935 GLK data, or to minimize it, as in the case of the 1936 RKB series. Thus, if the normal critical frequency at which the circuit is operating over the more northerly path is close to the cutoff frequency, the reduction in critical frequency will cause a complete fade-out over the more northerly path, while the critical frequency is still sufficiently

high to permit communication over the more southerly path. Furthermore, inasmuch as there is usually a decrease in absorption just before the critical frequency is reached, the lowering of the critical frequency may act to sustain, or actually increase, the fields over the southern path. If, however, the critical frequency happens initially to be so high that it is more than adequate to permit communication over both the northern and southern paths, the reduction in critical frequency may act advantageously on the critical frequency over the more northerly path to a greater degree than over the southern path. This condition will hold as long as the critical frequency does not fall below that necessary to maintain communication over the northern path, since, as already pointed out, the nearer the frequency of communication to the cutoff frequency, the less the absorption (up to the point where electron limitation begins to govern).

Conclusions

Results derivable from observations now available confirm the existence of a considerable latitude effect in the relationship of radio-transmission paths to magnetic disturbances.

For the period from January, 1935, to March, 1937, the magnitude of the observed differentials favoring the southern path during magnetic storms for the particular times and frequencies observed averages about 10 decibels for the British and Norwegian signals and about 6 decibels for the Russian signals. The combined average gives San Juan an advantage over Riverhead of approximately 8 decibels. The greater advantage gained in the case of the British and Norwegian signals, where the greater divergence in paths exists, over the Russian signals further confirms the actuality of the latitude effect.

Inclusion of data obtained for the severe magnetic storm of April, 1937, does not appreciably affect the results of the grand summary. The combined average advantage of San Juan over Riverhead is maintained at approximately 8 decibels. No data were taken on GLK and LCP during the April disturbance, but the average figures obtained for GLH and RKB remain substantially unchanged.

Considerable differences in the values obtained for individual storms exist. In one case the differential advantage is 21 decibels (GLK, storm of April 11, 1935). That the average advantage is smaller is probably due in part to the fact that the average reduction in the Riverhead level for all stations for all severe disturbances is in the order of 16 decibels. There is also a smoothing effect in the average, caused by the phenomenon of maximum reductions not always occurring on the same day at the two points of observation (see Fig. 6).

Several phenomena are of importance in determining the magnitude of the latitude effect. These are sufficiently complex to render a considerable series of observations desirable before a reasonably complete theory can be developed. It appears that both absorption and electron limitation are, in general, of importance in determining the magnitude of the changes observed, and that the relative importance of these factors are functions of frequency and time, as well as of transmission path. A complete consideration of all of these factors, therefore, requires a thorough appreciation of ionospheric conditions and their periodic changes over the paths under consideration. Data relating ionospheric conditions and latitude are as yet insufficient to permit a complete treatment of these relationships.

Acknowledgment

The authors wish to express their appreciation to the University of Puerto Rico for the facilities and cordial co-operation it accorded this research; to Mr. H. H. Beverage and Mr. H. O. Peterson, of RCA Communications, Inc., under whose direction the program was carried out; to Ruth H. Kenrick for assistance in the computation and compilation of the voluminous data; and to the many others who have assisted in the development of the program. Thanks are particularly due to Cable and Wireless, Ltd., of London, for arranging special transmissions and rendering valuable assistance in general, and to the Puerto Rican Light and Power Company, for assisting in the construction of the antenna at Rio Piedras.

Proceedings of the Institute of Radio Engineers
Volume 26, Number 7 *July, 1938*

A DIRECT-READING RADIO-WAVE-REFLECTION-TYPE ABSOLUTE ALTIMETER FOR AERONAUTICS*

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Summary—*Various theoretical and experimental researches have been carried out to indicate the absolute altitude of aircraft with respect to the ground by radio-wave-reflection methods. By using a new principle, frequency modulation, the following results have been obtained:*

1. *Completely continuous indication of altitude by a steady pointer and dial is easily accomplished and altitude variations occurring during time intervals of a few milliseconds are clearly detected.*
2. *Very small altitudes, less than four meters, are indicated quite easily.*
3. *The indication is quite linearly proportional to altitude.*
4. *The absolute error is zero and the relative error is less than a few per cent.*
5. *Altitudes in excess of 160 meters can be measured with a power consumption of only 3.9 watts.*

The above results show that this sounding gear is suitable for blind-landing use in aeronautics. Moreover, this gear may be applied to satisfy navigational requirements in foggy weather, to avoid the collision of ships, or to lead them into a port.

I. Introduction

THE ALTITUDE indicator for aeronautics generally used at present has the disadvantage that it cannot indicate absolute altitude with respect to the ground, since its principle is that of the aneroid barometer. It merely shows a relative altitude with respect to sea level or other initial reference level at constant (barometric) pressure. If the barometric pressure and the temperature vary, a large error, as great as several hundred meters, may occur. Hence, it is not entirely suitable for blind-landing use, for which an accurate instrument is urgently needed, and does not reduce the great inconvenience and danger in low-altitude flying. To eliminate these disadvantages several methods have been proposed, such as the use of a light beam or acoustic echo, etc. However, these methods are not in practical use because of various disadvantages.

An absolute altimeter utilizing a radio-wave-reflection method, has the following outstanding advantages:

1. No acoustic interference from the airplane engine.
2. No fading phenomena caused by wind as in the method utilizing a sound wave.

* Decimal classification: R526.4. Original manuscript received by the Institute, November 29, 1937.

3. Since the attenuation of a radio wave is small, the maximum measurable altitude should be in excess of 10,000 meters.
4. An altitude of several kilometers can be measured with low power.

An absolute altimeter may be based on any of several principles, among which are:

A. Electric-Capacitance Principle

This principle utilizes the capacitance between an antenna stretched on the airplane and the earth which varies with altitude. However, this capacitance variation is relatively large only for quite low altitudes, but is very small for high altitudes. Therefore, it is not considered practicable.

B. Phase-Discrimination Principle

There is a phase difference proportional to altitude, between a wave radiated from an airplane and its reflection. By its measurement, the altitude can be determined. However, this is very difficult and this principle is considered improbable of realization in a practical instrument.

C. Radio-Wave-Impulse Principle

This utilizes the method often used to measure the ionosphere height; namely, a radio wave of very short duration is transmitted and the echo time observed by means of a cathode-ray tube. This is almost impossible at the present state of the art for the following reasons:

1. If the altitude is 10 meters; it is necessary to produce a radio-wave impulse shorter than $1/15\times10^6$ second. The shortest impulse produced at present is of about $1/10^4$ second duration. Therefore, this method is utterly eliminated from consideration.

2. Even though such a very short impulse may be produced further difficulties prevent realization of the objective. For instance, harmonic-wave analysis shows that such an impulse contains a band of ultra-high frequencies which is several times ten megacycles in width. To amplify faithfully the very short impulse received requires an amplifier having a very wide frequency band. To obtain such an amplifier is evidently difficult.

Many years of theoretical and experimental researches have resulted in the development of a frequency-modulation principle, on which is based a direct-reading absolute altimeter suitable for blind-landing use.

II. Principle

A frequency-modulated radio wave is radiated from the plane. The beat frequency resulting from the combining of the radiated wave and the reflected one from the ground is a function of the altitude. A con-

tinuously operating meter indicating the beat frequency may be calibrated in absolute altitude.

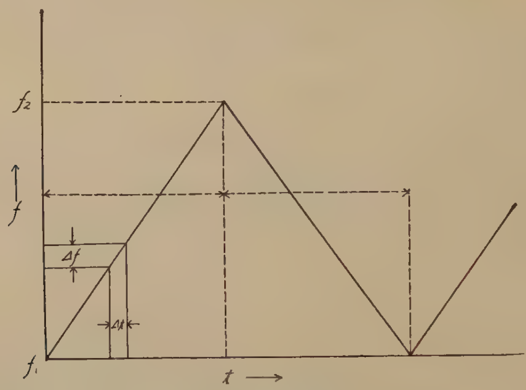

Fig. 1—Frequency characteristic of radio emitter.

The frequency characteristic of the emitted wave should be linearly proportional to time t, as is shown in Fig. 1. The ordinate scale shows the frequency f. The beat frequency between the radio wave emitted

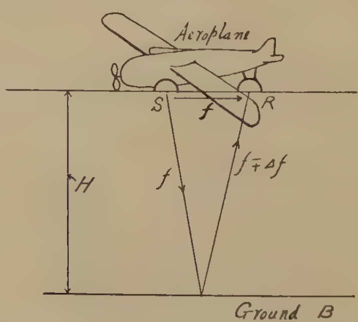

Fig. 2—Principle of radio altimeter.

from S (Fig. 2) and the reflected one from the ground B should be Δf. Between these waves there is a time difference Δt proportional to altitude H,

$$\Delta t \propto 2H.$$

From Fig. 2,

$$\Delta f \propto \Delta t$$

and,

$$\Delta f \propto 2H. \tag{1}$$

Therefore Δf is directly proportional to the altitude H. If Δf is measured by an indicating-type frequency meter, the indication will be proportional to the altitude.

III. Determination of Carrier Frequency

Theoretically, the carrier frequency may be of any value, but if the wavelength is made short, the following advantages are obtained:

1. High directivity.
2. Light weight and small size of equipment which is most important in aeronautical work.
3. Frequency variation, i.e., $f_2 - f_1 = fd$ is comparatively large; thus, a very low altitude may be measured.

For the above reasons a wavelength of about 50 centimeters, corresponding to a frequency of about 6×10^8 cycles was used.

IV. Radio-Wave Emitter

To emit radio waves of the above frequency, back-coupled, electron, or magnetron oscillators, etc., may be used. An electron oscillator

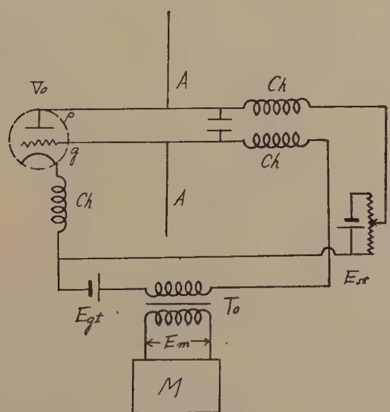

Fig. 3—Emitter connection diagram.

was preferred because of its ease of handling. By varying its grid voltage, sufficient frequency modulation was obtained.

A. Oscillator

The circuit diagram is shown as Fig. 3. V_0, A, and M show, respectively, the oscillator valve, antenna, and modulator valve.

If positive voltage is applied to the grid g and negative voltage to the plate p of the oscillator valve V_0, and they are adjusted properly, electron oscillations of the Barkhausen-Kurz type will be produced. The frequency f is about

$$f \propto \sqrt{E_{gt}} \tag{2}$$

where E_{gt} is the grid voltage.

Now suppose that an alternating current, with a wave form as shown in Fig. 1, is applied to the grid from modulator M. The desired frequency modulation will be obtained by adjusting the alternating-current amplitude E_m. In case the value of E_m is very small compared with E_{gt}, (2) is approximately:

$$f \propto E_{gt}, \tag{3}$$

or, the frequency variation is approximately proportional to the total grid voltage.

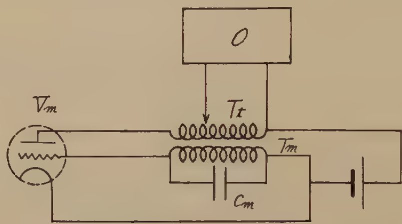

Fig. 4—Modulator connection diagram.

The directivity characteristics as calculated for the antenna are as follows:

divergence angle	$\alpha = 27°$
effective angle	$\gamma = 18°$
directivity efficiency	$\eta = 67$

B. Modulator

The circuit diagram is shown as Fig. 4. V_m, T_m, C_m show, respectively, the modulating oscillator valve, coil, and condenser. By adjusting tap T_t, the desired voltage is applied to the centimeter-wave oscillator O.

V. Receiver

The receiver contains a detector, amplifier, and limiting output stage.

A. Detector

Fig. 5 gives the circuit diagram of the detector. V_γ, A_γ, A_m, C_h, C_γ, T_γ are, respectively, the detector valve, receiving antenna, amplifier, choke coil, by-pass variable condenser, and the audio-frequency transformer. By properly adjusting the grid voltage $E_{g\gamma}$, the plate voltage $E_{p\gamma}$, and the by-pass condenser C_γ, ultra-high-frequency radio waves may be detected.[1]

[1] Uda, "A new type receiving set for extremely short waves." *Jour. I.E.E.* (Japan), vol. 49, p. 724; June, (1929).

B. Amplifier

A resistance-capacitance-coupled amplifier was adopted because of the necessity for amplifying low frequencies and its good response-

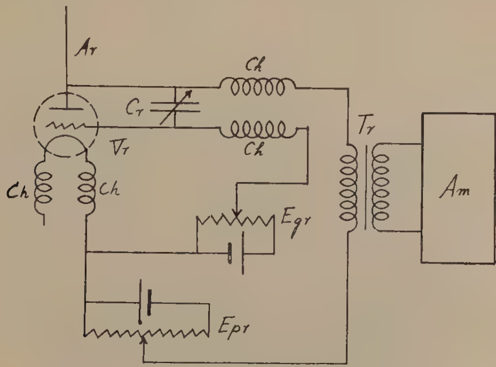

Fig. 5—Detector connection diagram.

frequency characteristic. The amplifying valves used are two high-frequency pentodes. Its actual voltage amplification is about 40,000.

C. Limiting Output Stage.

Although the reflected-wave amplitude is inversely proportional to the square of the height to be measured, the beat-frequency voltage should be limited to a constant level. For this limiter valve, a heater-type triode is used and is worked also as the output valve.

VI. Altitude Indicator

The beat frequency of constant output level will be measured by any indicator-type frequency meter. After various trials, the one found most appropriate is as follows. Fig. 6 shows the principle.

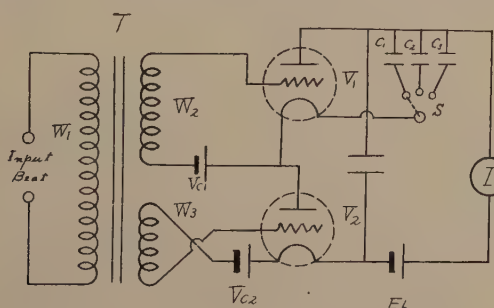

Fig. 6—Altitude indicator circuit diagram.

T is a transformer with three windings, W_1, W_2, and W_3. The secondary windings, W_2 and W_3, are connected to give inverse voltage polarity to triodes V_1 and V_2. Plate voltage E_b and grid bias V_{C_1} and

V_{C_2} are so adjusted that no current flows in the milliammeter I when no input beat voltage is applied to the primary winding W_1. When a beat voltage Δf is applied to winding W_1, V_1 and V_2 alternately become conductive and nonconductive. During the moment when V_1 is nonconductive and V_2 is conductive, the capacitance C_1 is charged by E_b. Its charge Q is

$$Q = C_1 E_b.$$

During the moment when V_1 is a conductor and V_2 a nonconductor,

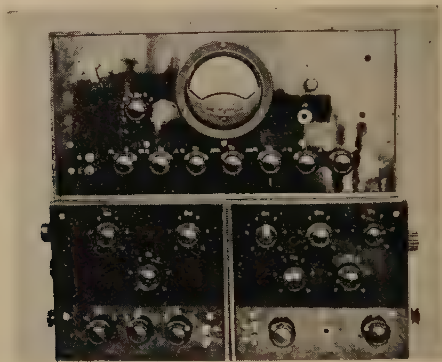

Fig. 7—Direct-reading radio-wave absolute altimeter, outside view.

this charge is discharged through V_1. This action is then repeated Δf times per second; thus the milliammeter current I is

$$I = \Delta f Q = \Delta f C_1 E_b.$$

If C_1 and E_b are held at constant values in the above equation,

$$I = K \Delta f. \tag{4}$$

That is, the input beat frequency should be indicated directly. As shown by (1) and (4), it is evident that this altimeter indication I is directly proportional to altitude. Therefore completely continuous altitude indication by a steady pointer and dial is easily accomplished.

This altitude indicator was most adequate for this purpose because of its high input impedance, low power consumption, and constant indication even though the input voltage varies from a few volts to several hundred volts, which is not characteristic of many other indicating-type frequency meters. By switching to either C_1, C_2, or C_3, the highest accuracy may be obtained for the various altitudes of the airplane.

VII. Experimental Sets

Fig. 7 shows the outside view of an experimental set. The radio-wave emitter is shown at the lower left, the receiver at the lower right,

and the amplifier, limiter, and altitude indicator in the upper section. Fig. 8 shows an inside view of the radio-wave emitter (right) and receiver (left). Fig. 9 shows an interior view of the amplifier, limiter, and altitude indicator.

Fig. 8—Emitter, inside view.

Fig. 9—Amplifier, limiter, and altitude indicator, interior view.

VIII. Experiment for Altitudes of Less than 50 Meters

A. Method

This experiment was carried out at the football field of Tohoku Imperial University. Considering the metallic net of the field as equivalent to the earth, the distance between it and the apparatus was varied and the altitude-indicator-current recorded. Fig. 10 shows the ground and the metallic net Z.

The metallic net is made of iron wire of 7-centimeter square mesh and its dimensions are 10×30 meters.

B. Radio-Wave Emitter

1. Oscillator

The grid input of the oscillator, described in Section IVA, is 22 milliamperes at 115 volts or 2.5 watts. During frequency modulation the

total frequency deviation from maximum to minimum is 38 megacycles.

2. Modulator

The plate input of the modulator, described in Section IVB, is 25 milliamperes at 150 volts. The modulator frequency is 25 cycles.

Fig. 10—Field of experiment.

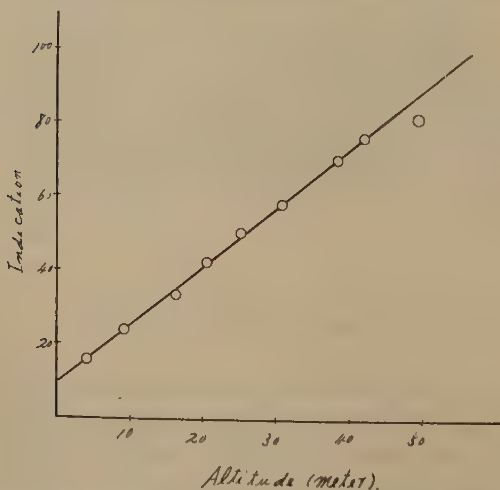

Fig. 11—Experimental results for altitude of less than 50 meters.

C. Receiver

1. Detector

The grid input of the detector, described in Section VA, is 3 milliamperes at 140 volts. On the plate, a bias of 1.8 volts is applied.

D. Results

The results will be seen in Fig. 11.

As shown in these, the indications are approximately linearly proportional to altitude and prove the former theory distinctly. The minimum altitude indicated is 4 meters.

IX. Experiment for Altitudes of More Than 50 Meters

A. Method

This experiment was carried out by the same method described in Section VIIIA.

B. Radio-Wave Emitter

1. Oscillator

The grid input of the oscillator, described in Section IVA, is 26 milliamperes at 150 volts or 3.9 watts. Its total frequency deviation is 20 megacycles.

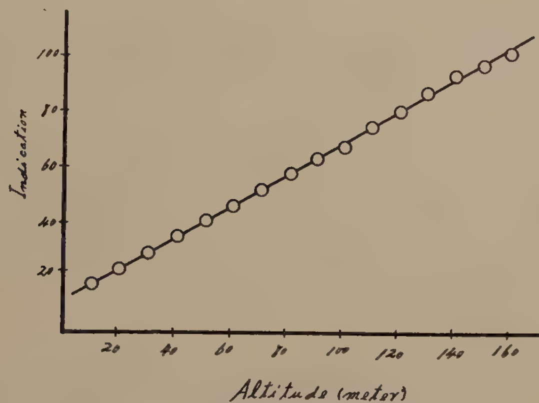

Fig. 12—Experimental results for altitude of more than 50 meters.

2. Modulator

The plate input of the modulator, described in Section IVA, is 25 milliamperes at 150 volts. The modulation frequency is 24 cycles.

C. Receiver

1. Detector

The grid input of the detector, described in Section VA, is 2.5 milliamperes at 200 volts, and its plate current is 25 microamperes.

D. Results

The results will be seen in Fig. 12. As shown, the indications are linearly proportional to altitude.

In this case altitudes greater than 160 meters could not be measured because that was the maximum dimension of the ground.

X. Conclusion

By using the new principle of frequency-modulation the following results have been obtained:

1. Completely continuous altitude indication by a steady pointer and dial is accomplished, altitude variations occurring during a few milliseconds are clearly detected.
2. A very small altitude, less than 4 meters, is indicated quite easily.
3. The indication is quite linearly proportional to the altitude.
4. The absolute error is zero and the relative error is less than a few per cent.
5. With a transmitter power consumption of only 3.9 watts, altitudes of 160 meters or more may be measured.

The above results show that this sounding gear is suitable for blind-landing use in aeronautics.

Moreover, this gear may be applied to satisfy navigational requirements in foggy weather, to avoid collision of ships, or to lead them into a port.

XI. Acknowledgment

This research has been carried on in the Tohoku Imperial University, Sendaishi, and has been aided by financial support from the Invention Bounty of the Commercial and Industrial Ministry, the Oi Fund of the Institute of Electrical Engineers of Japan, and the Saito Gratitude Foundation.

Dr. Honda, President of the Tohoku Imperial University, and Dr. Mitsuda, Chief of the Electrical Research Laboratory of the Communicational Ministry, gave me effective suggestions. I am glad to have this opportunity of expressing my appreciation of this generous support and help.

Proceedings of the Institute of Radio Engineers
Volume 26, Number 7 *July, 1938*

THEORY OF THE DIODE VOLTMETER*

By
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Summary—*The current-voltage characteristic of a diode can be expressed at low currents by an equation of the form $i = G\epsilon^{kE}$. An analysis based on this relation shows that the rectification efficiency and input resistance of the usual form of grid-leak-condenser diode circuit can be conveniently expressed entirely in terms of the grid-leak resistance R_1, the amplitude of the impressed alternating voltage E, a single parameter k of the tube, and an adjustable constant b. k is determined by the cathode temperature, and is constant, for a given tube, as long as the heater voltage is unchanged. b is equal to k times the voltage which exists across R_1 in the absence of an impressed alternating voltage, and can be adjusted to any value within a wide range by the use of a suitable bias battery.*

It is shown that large values of b give higher efficiencies of rectification, but that, if b is of the order of 100 or more, the efficiency is practically as high as can be obtained, and is nearly independent of b.

The various components of alternating current flowing through the diode are given by $A_n = 2A_0 f_n(kE)$, in which A_n is the amplitude of the nth harmonic of the impressed frequency and A_0 is the direct current through the diode, which last can be separately computed. $f_n = I_n(kE)/I_0(kE)$. $I_n(x)$ is the nth order Bessel's function of a pure imaginary argument.

The input resistance R_i of the diode when connected with its grid leak and condenser across a tuned circuit is given by

$$\frac{R_i}{R_1} = \frac{x}{2f_1(x)(\beta x + b)},$$

in which $x = kE$ and β is the efficiency of rectification. For very small impressed voltages, this reduces to

$$R_i = \frac{R_1}{b} = \frac{1}{kA_0'},$$

in which A_0' is the direct current through R_1 in the absence of a signal. Curves showing the variation of input resistance with signal amplitude are given for three values of b. These show that if b is small the input resistance will decrease with increasing signal, while if b is large the opposite is true.

Finally, an analysis is made of the effect of an internal resistance in the circuit to which the diode voltmeter is connected, this resistance being independent of frequency. A formula is derived which gives the value which this resistance must have to bring about a 1 per cent reduction in the rectified voltage. This formula is expressed in terms of b, x, R_1, and β, the last being itself a function of x which can be obtained from curves that are given.

* Decimal classification: R134×R243.1. Original manuscript received by the Institute, January 17, 1938. Presented before joint meeting of I.R.E. and A.A.A.S., Indianapolis, Indiana, December 28, 1937.

List of Symbols

C, R_0, and R_1 = circuit elements located as shown in Fig. 1.
E_c = bias battery voltage.
E = amplitude of the impressed sinusoidal electromotive force.
i = total current through the diode.
A_0 = direct-current component of current through the diode.
A_n = amplitude of component of current through diode of frequency. $n\omega/2\pi$.
A_0' = value of A_0 when $E=0$.
$E_1' = R_1 A_0'$ = voltage across R_1 when $E=0$.
$\Delta E_1 = R_1(A_0 - A_0')$ = rectified voltage.
$\beta = \Delta E_1/E$ = efficiency of rectification.
G and k = parameters of the diode.
$x = kE$.
$b = kE_1' = kR_1A_1'$.
R_i = input resistance of the diode.
R_i' = input resistance of the diode when $E=0$.
$J_n(x)$ = the nth order Bessel function of x.
$I_n(x) = \frac{1}{(j)n} J_n(jx)$ = another kind of nth order Bessel function. It is a real quantity.
$f_n(x) = I_n(x)/I_0(x)$.
a_0 = change in A_0 caused by inserting resistance at R_0 in Fig. 8.
$B_0 = A_0 + a_0$.

$$N = a_0/\Delta A_0 = \frac{a_0}{A_0 - A_0'}.$$

$$F(x) = \frac{I_0(2x)}{I_0^2(x)} - 1.$$

It has long been recognized that, at low currents, an exponential relation of the form

$$i = G\epsilon^{ke} \tag{1}$$

exists between the current through a high-vacuum thermionic diode and the voltage impressed across its terminals. Twelve years ago, Colebrook[1] published an analysis of the conventional diode detector circuit of Fig. 1, in which it was assumed that the rectifier obeyed this law. The developments which he obtained are interesting and important, but his results are expressed in the form of infinite series and his attention was confined to rather small signals, so that it was not evi-

[1] F. M. Colebrook, "The rectification of small radio frequency potential differences by means of triode valves," *Exp. Wireless and Wireless Eng.*, vol. 2, November and December, (1925).

dent just how useful and convenient the analysis of the exponential conductor can be in dealing with the diode. Later on, other authors[2,3] developed more compact expressions for the rectified voltage delivered

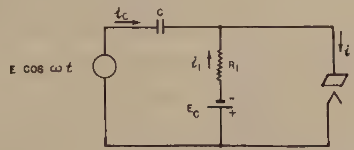

Fig. 1—The simple diode voltmeter connected to an alternating-current generator of negligible internal impedance.

by an exponential diode, but did not otherwise extend Colebrook's work on this most common of detector circuits.

It is the purpose of this paper to develop more fully the circuit analysis and to describe the behavior and peculiarities of the typical diode voltmeter. The theory is also applicable to the detector of the ordinary radio-receiving set in so far as its response to the carrier wave goes. The analysis is not, however, extended here to the consideration of modulated waves.

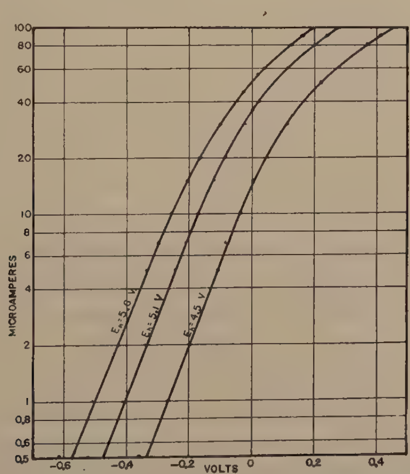

Fig. 2—Current-voltage characteristics of the diode elements in a 6B7. The two auxiliary anodes are connected in parallel.

Characteristics of Typical Diodes

Fig. 2 shows semilogarithmic plots of the current-voltage characteristics of the two diode elements of a 6B7 tube connected in

[2] Groeneveld, Posthumus, and van der Pol, "Gittergleichrichtung," *Zeit. für Hochfrequenz.*, vol. 29, pp. 139–147; May, (1927).
[3] M. J. O. Strutt, "Gleichrichtung," *Hochfrequenz. und Electroakustik*, vol. 42, pp. 206–208; December, (1933).

parallel, while Fig. 3 shows similar curves for a 6H6. The former are straight lines up to about 10 microamperes and the latter to about 20 microamperes. The 6B7 curves bend over fairly rapidly for higher currents, while the 6H6 curves bend more gradually. It has been shown theoretically[2] that the slope of the straight portion of the characteristic of a given tube should be inversely proportional to absolute temperature, and it may be seen that the curves for lower heater voltage are steeper than those for the higher voltages, although the difference is not very great.

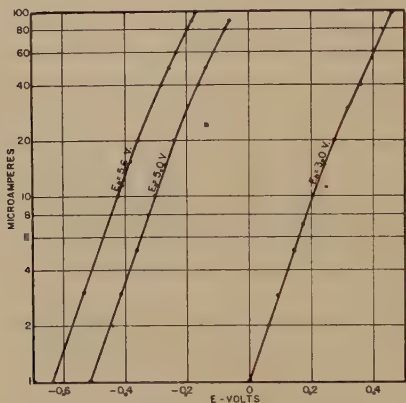

Fig. 3—Current-voltage characteristics of a type-6H6 tube. The two diode elements are connected in parallel.

The curve for $E_A = 5.6$ volts was studied below 1 microampere in the case of the 6H6, and it appears to depart from the straight-line relation below 0.5 microampere, due, apparently, to a trace of ionization current in the tube. Where such currents are absent, it seems likely that the exponential law will be valid at very low current levels. With this tube, it is obeyed over a current range of 40 to 1 at 5.6 heater voltage, and probably over a wider range at lower heater voltages, since the ionization current was then smaller.

The value of k which appears in (1) can be determined from the slope of these curves, but is better obtained by the use of a dynamic method. This involves biasing the diode to some convenient small current, such as 2 microamperes, and then applying a sinusoidal voltage of just the right magnitude to double the current reading. k is then given by

$$k = 1.81/E_a, \tag{2}$$

in which E is the amplitude of the impressed alternating voltage. This

formula has been justified in another paper.[4] Since k is inversely proportional to temperature, it is higher for oxide-coated than it is for thoriated or tungsten cathodes. As it is the oxide type which is most likely to be used and as a high value of k will be shown to be very advantageous, we need not consider tubes having the other types of cathode. For the oxide type, k will ordinarily lie between 8 and 14, and, if an extremely low cathode temperature is not used, we may take it to be between 8 and 12. Values of k within this range have been given for certain European tubes by Groeneveld.[2] If the cathode temperature is held at a fixed value, k is reasonably constant.

The parameter G is of less importance. It is the current which flows with zero impressed voltage, and its magnitude will depend upon the cathode temperature. However, the current which flows in the absence of an impressed alternating voltage can be arbitrarily adjusted by the use of a bias voltage, and the constant G in (1) will then be replaced by $G\epsilon^{-kE_c}$, the positive pole of E_c being connected to the cathode. This product is a new constant and can be given any desired value within a range determined by the limits of validity of the exponential relation (1). We need, therefore, give no further consideration to G.

Applicability of the Theory of the Exponential Conductor to the Circuit of Fig. 1

It might seem at first glance that, since an exponential law is obeyed over only a rather limited range of impressed voltage, which varies with different tubes from about 0.4 to 1.0 volt, the use of this relation would not greatly extend the conventional "small-signal" theory of rectifier action. This is by no means the case, as a moment's reflection will show. The essence of the small-signal theory lies in expressing the rectifier current in the form

$$i = a_0 + a_1 e + a_2 e^2. \tag{3}$$

Such an equation can be made to represent with some accuracy a limited portion of the current-voltage characteristic just to the right of cutoff, and the rectification of small signals applied to this portion of the characteristic can be easily calculated. If the circuit of Fig. 1 is used and a large voltage is applied, the diode is automatically biased so that only the positive crests of the impressed wave pass to the right of cutoff. These crests will fall upon the part of the characteristic which is represented by (3), and if the equation gave zero current to the left of cutoff, it could be used to predict, with moderate accuracy, the per-

[4] C. B. Aiken, "Sharp cut-off in vacuum tubes with applications to the slide-back voltmeter." *Elec. Eng.*, vol. 57, pp. 171–176; April, (1938).

formance of the circuit. This, however, is not the case. It is quite evident that (3) gives a minimum for a certain value of e, and increasing currents as e is made more negative. An equation of this form is therefore strictly limited to small signals.

On the other hand, if we express the characteristic as in (1), we may choose the parameters G and k so as to fit accurately the lower portion of an experimentally determined curve, and at large negative values of voltage the current will become negligibly small. Whether the theoretical and actual curves agree at very low values of current is of little practical importance, since it is the more markedly conducting range that will determine the behavior of the rectifier. This law is therefore well suited to a study of the detector circuit of Fig. 1, and will give accurate results, provided the external resistance R_1 is not made so small that large currents will flow. If this is allowed to happen, the operation will include the upper portions of the current-voltage characteristic, in which the exponential law no longer holds. However, small values of R_1 are not ordinarily used in practice, and hence this limitation is of little consequence.

In experimental studies of the diode circuit, it is convenient to measure the continuous voltage across R_1 with a direct-current vacuum-tube voltmeter, as the current through R_1 is often too small to be conveniently read with a microammeter.

Analysis of the Exponential Diode

Some of the theory which immediately follows has been given by Groeneveld, and a certain amount of repetition is ventured for the sake of clarity and completeness. It will be assumed that the reactance of the condenser C is negligible at the frequency of the impressed voltage. In the first part of the paper, it will also be assumed that the internal impedance of the generator is negligible, the effect of such impedance being considered in a later section. Let us assume that the impressed voltage is $E \cos \omega t$, and let A_0 be the continuous component of current through the rectifier. The amplitude of the fundamental frequency component of current through the rectifier will be A_1, that of the second harmonic, A_2, etc. Because of the presence of C, the continuous current through R_1 will also be A_0, and the only voltages appearing across the diode will be the impressed alternating voltage and the voltage due to the sum of E_c and the potential drop across R_1, or

$$e = E \cos \omega t - E_c - R_1 A_0. \tag{4}$$

Inserting this into (1), we obtain

$$i = G\epsilon^{-kE_c}\epsilon^{-kR_1A_0}\epsilon^{kE\cos\omega t}. \tag{5}$$

Expressing i in terms of its components, (5) can be written

$$\epsilon^{kR_1A_0}[A_0 + A_1 \cos(\omega t + \theta_1) + A_2 \cos(2\omega t + \theta_2) + \cdots] = G\epsilon^{-kE_c}\epsilon^{kE\cos\omega t}. \tag{6}$$

It is well known that the quantity $\epsilon^{kE\cos\omega t}$ can be expressed in a Fourier series of the form

$$\epsilon^{kE\cos\omega t} = I_0(kE) + 2\sum_{n=1}^{\infty} I_n(kE)\cos n\omega t, \tag{7}$$

which, substituted into (5), gives

$$\epsilon^{kR_1A_0}[A_0 + A_1\cos\omega t + A_2\cos 2\omega t) + \cdots] = G\epsilon^{-kE_c}[I_0(kE) + 2I_1(kE)\cos\omega t + 2I_2(kE)\cos 2\omega t + \cdots], \tag{8}$$

it being evident by inspection that the values of θ_1, θ_2, etc., must be zero. In these equations, $I_n(kE)$ is a particular variety of nth-order Bessel's function. It is a real quantity and is related to the ordinary Bessel's function, which is usually represented by J_n, in the following manner:

$$J_n(jx) = (j)^n I_n(x). \qquad j = \sqrt{-1}. \tag{9}$$

Tables of $I_0(x)\epsilon^{-x}$ and $I_1(x)e^{-x}$ for values of x up to 16 are to be found in Watson's book.[5] Tables of $I_0(x)$ and $I_1(x)$ are to be found in Jahnke and Emde[6] for values of x from 0 to 9, as are also tables of $I_n(x)$ for $x=0$ to $x=6$ and of n from 0 to 11. For x large, the following series expressions are valid:

$$I_0(x) = \frac{\epsilon^x}{\sqrt{2\pi x}}\left[1 + \frac{1}{8x} + \frac{9}{128x^2} + \cdots\right] \tag{10}$$

$$I_1(x) = \frac{\epsilon^x}{\sqrt{2\pi x}}\left[1 - \frac{3}{8x} - \frac{15}{128x^2} \cdots\right] \tag{11}$$

$$I_2(x) = \frac{\epsilon^x}{\sqrt{2\pi x}}\left[1 - \frac{15}{8x} + \frac{105}{128x^2} \cdots\right] \tag{12}$$

$$I_3(x) = \frac{\epsilon^x}{\sqrt{2\pi x}}\left[1 - \frac{35}{8x} + \frac{910}{128x^2} \cdots\right]. \tag{13}$$

Efficiency of Rectification

The direct-current terms in (8) can be separately equated, and give

$$A_0\epsilon^{kR_1A_0} = G\epsilon^{-kE_c}I_0(kE). \tag{14}$$

[5] G. N. Watson, "Theory of Bessel Functions," Cambridge University Press, (1922).

[6] Jahnke and Emde, "Tables of Functions," second edition. B. G. Teubner, Berlin, (1933).

Let us represent by A_0' the value which A_0 has when E is zero.

$$A_0'\epsilon^{kR_1A_0'} = G\epsilon^{-kE_c}, \tag{15}$$

since $I_0(0)=1$.

The rectified voltage is equal to the change in the voltage drop across R_1 which results when the alternating voltage E cos ωt is impressed. Representing this change by ΔE_1, it is evident from (14) and (15) that

$$\Delta E_1 = R_1(A_0 - A_0') = R_1G\epsilon^{-kE_c}[I_0(kE)\epsilon^{-kR_1A_0} - \epsilon^{-kR_1A_0'}] \tag{16}$$

$$\Delta E_1 = R_1G\epsilon^{-kE_c}\epsilon^{-kR_1A_0'}[I_0(kE)\epsilon^{-kR_1(A_0-A_0')} - 1], \tag{16a}$$

which, in view of (15), gives

$$\Delta E_1 = R_1A_0'[I_0(kE)\epsilon^{-k\Delta E_1} - 1]. \tag{17}$$

Now, let $R_1A_0'=E_1'$ and $kE_1'=b$. Then

$$(\Delta E_1 + E_1')\epsilon^{k\Delta E_1} = E_1'I_0(kE), \tag{18}$$

and, multiplying both sides by k, there results

$$\epsilon^{-k\Delta E_1}(k\Delta E_1 + b) = bI_0(kE). \tag{19}$$

This equation shows that the rectified voltage is a function only of the impressed-voltage amplitude E, the tube parameter k, and the quantity b. This last is k times the voltage E_1' which appears across R_1 in the absence of an impressed alternating-current wave. E_1' is determined entirely by E_c, R_1, and the parameters G and k, as is evident from (15). In fact, (15) may, for convenience, be rewritten

$$E_1'\epsilon^{kE_1'} = R_1G\epsilon^{-kE_c}, \tag{20}$$

or, in terms of b,

$$b\epsilon^b = kR_1G\epsilon^{-kE_c}. \tag{21}$$

An ordinary diode operated without bias will have a value for E_1' of several tenths of a volt if reasonably large values of R_1 are used. k may be taken between 8 and 12 for oxide-coated tubes, and consequently b will ordinarily lie between about 4 and 14 if no bias battery is employed. It will appear presently that large values of b increase the efficiency of rectification, and therefore a bias may be used intentionally to raise A_0' and thereby increase b.

Let us define the efficiency of rectification as $\Delta E_1/E$ and represent it by β. By assigning values of $k\Delta E_1$ in (19), we can solve for $I_0(kE)$. kE is most readily determined by plotting $I_0(x)$ versus x on semilog paper and working backwards from the calculated value of $I_0(kE)$.

Curves of β versus $x=kE$ are shown in Fig. 4. If $b=\infty$, the curve is indistinguishable from that of $b=100$, except at the right end, where the difference in β for the two cases is less than 0.01. It will be observed that the efficiency of rectification does not change greatly with b, if b is large. From these curves, we may readily predetermine the rectified voltage which will result from any applied alternating electromotive force, provided that the parameter k is known and that E_1' is approximately known. E_1' is easily determined by measuring the initial current through the tube and multiplying by R_1, or by direct measurement with a direct-current vacuum-tube voltmeter. As has been pointed out, k is constant for a given tube if the operating temperature is fixed.

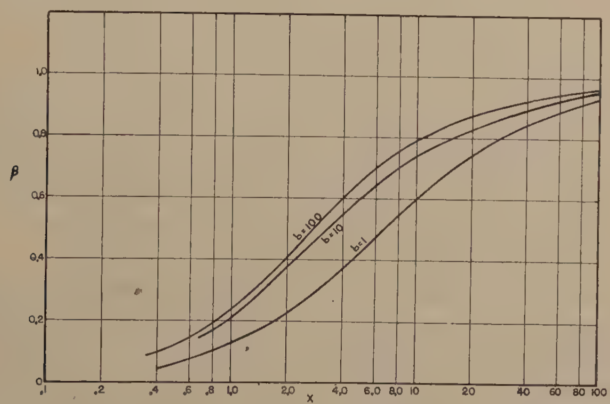

Fig. 4—Efficiency of rectification as a function of $x=kE$. The parameter $b=kR_1$ times the initial direct current through the diode.

Certain conclusions may be drawn from these curves. First of all, it is evident that the larger k is, the more efficient the rectification for a given impressed voltage will be, since a bigger k will give a larger value of x. If b is of the order of 8, as is likely to be the case, it is possible to obtain some improvement in rectifying efficiency by employing a positive bias of a few volts. If this is done, a large value of R_1, say 5 megohms or more, should be used in order that the initial current may be kept small, which will insure operation on the exponential portion of the diode characteristic. Departure from this portion will lower the efficiency.

Experimental checks of the equivalent of (17) for the rectified voltage have been reported by Colebrook[1] and Groeneveld[2] for impressed voltages of 1 volt peak and less. However, it appears that this formula, and its more convenient form (19), should be valid for very much larger voltages if R_1 is not too small, and consequently

some measurements were made to establish this fact. Fig. 5 shows theoretical curves which are calculated from those of Fig. 4, while the dots show the experimentally determined points. k for the 6H6 tube used was found by the dynamic method of measurement to be 11.6, the heater voltage being 5.6 volts. The values of R_1 were 0.1 and 5.22 meghoms for the lower and upper curves, respectively, and a bias voltage was used in each case to adjust b to the value indicated.

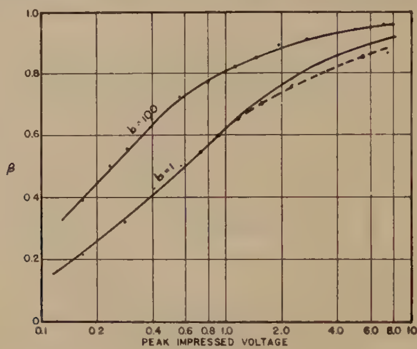

Fig. 5—Experimental checks of the curves of Fig. 4.

The check between theory and experiment is very close throughout the length of the upper curve, as it is in the left-hand portion of the lower one. The departure of the measured points from the right-hand end of the theoretical curve for $b=1$ is not surprising, since, with so low a value of R_1, the current swing is large. When $b=100$, there is no discrepancy for the largest measured value of 7.07 volts peak, and since the rectification efficiency is 0.95, we should expect no great divergence between measured and calculated points for larger impressed voltages, as we are already within 5 per cent of the maximum possible value of β. This curve shows clearly that the theory of the exponential conductor can be applied to the diode voltmeter, even when large voltages are impressed. Furthermore, it will give fairly accurate results for the ordinary detector, in which R_1 is between 0.25 and 1.0 megohm. In both cases, a far greater range of voltages may be dealt with than can be handled by small-signal theory.

Input Resistance of a Diode Voltmeter Connected to a Tuned Circuit

When the detector of Fig. 1 is connected across a tuned circuit in which flows an unmodulated current, the only alternating voltage impressed across the diode is that having a frequency to which the tuned circuit is resonant, and the theory which has been developed

applies without modification. We must, however, develop an expression for the alternating current through the diode, and this can be readily done by an examination of (8). It is evident from inspection of this equation that

$$A_1 = G\epsilon^{-kE}\epsilon^{-kR_1A_0}I_0(kE)\left[\frac{2I_1(kE)}{I_0(kE)}\right]. \tag{22}$$

Let

$$f_1(kE) = f_1(x) = \frac{I_1(kE)}{I_0(kE)}. \tag{23}$$

Then, by (15) and (23), (22) may be written

$$A_1 = 2A_0 f_1(kE). \tag{24}$$

A_0 is the total continuous component of current through the rectifier and is itself a function of the impressed voltage E. It may, if desired, be readily calculated as follows:

$$R_1A_0 = \Delta E_1 + E_1' \tag{25}$$

$$\frac{R_1A_0}{E} = \frac{\Delta E_1}{E} + \frac{kE_1'}{kE} = \beta + \frac{b}{x}, \tag{26}$$

or,

$$kR_1A_0 = k\Delta E_1 + kE_1' = \beta x + b. \tag{26a}$$

The ratio of R_1A_0 to E can be calculated from this expression by making use of the curves of Fig. 4.

Again inspecting (8), we have

$$A_2 = 2A_0 f_2(kE) \tag{27}$$

$$A_3 = 2A_0 f_3(kE) \tag{28}$$

$$\cdots\cdots\cdots$$

$$A_n = 2A_0 f_n(kE), \tag{29}$$

in which

$$f_n(kE) = \frac{I_n(kE)}{I_0(kE)}. \tag{30}$$

Plots of $f_1(x)$, $f_2(x)$, and $f_3(x)$ are given in Fig. 6. Evidently, when $x=100$, the second and third harmonics of diode current are nearly as large as the fundamental.

The input resistance of the diode is E/A_1, and, by (24), this is

$$R_i = \frac{E}{2A_0 f_1(kE)}, \tag{31}$$

or,

$$\frac{R_i}{R_1} = \frac{1}{2f_1(x)\dfrac{R_1}{E}(A_0 - A_0' + A_0')}$$

$$= \frac{1}{2f_1(x)\left[\dfrac{R_1(A_0 - A_0')}{E} + \dfrac{kR_1A_0'}{kE}\right]}$$

$$\frac{R_i}{R_1} = \frac{1}{2f_1(x)\left(\beta + \dfrac{b}{x}\right)}, \tag{32}$$

since $R_1(A_0 - A_0')$ is the rectified voltage. β can be obtained from Fig. 4 for any value of x and $f_1(x)$ from Fig. 6.

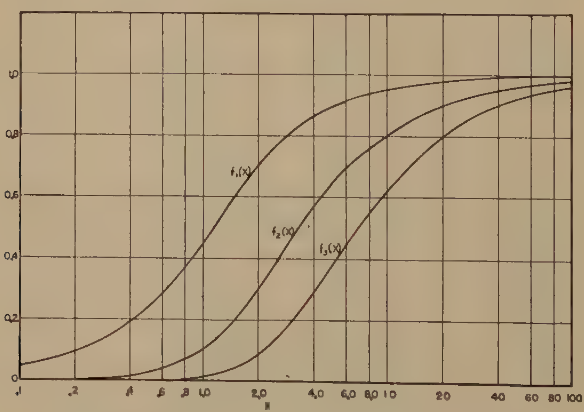

Fig. 6—Plots of the functions defined by (30).

Equation (32) gives us in simple form the ratio of the resistance of the diode to the external resistance R_1. It is interesting to compare this formula with the expression for R_i/R_1 frequently given[7] for the diode detector, which is

$$\frac{R_i}{R_1} = \frac{1}{2\beta}. \tag{33}$$

[7] F. E. Terman, "Radio Engineering," second edition, p. 426. McGraw-Hill Book Company, (1937).

When x is very large, (32) approaches (33) as a limit, since $f_1(x)$ approaches unity. In this range β is also fairly close to unity, and both (32) and (33) are near the value 0.5. For smaller values of x, there is a considerable difference between the two expressions, and when x is quite small (33) becomes very large, while (32) becomes

$$\frac{R_i}{R_1} \to \frac{1}{x\left(\beta + \dfrac{b}{x}\right)} \to \frac{1}{b} = \frac{R_i'}{R_1}, \qquad (34)$$

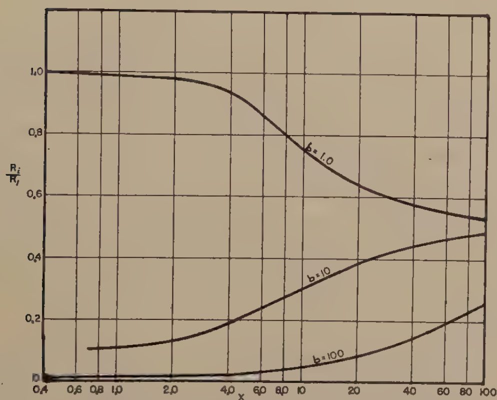

Fig. 7—Effective input resistance of a diode voltmeter connected across a tuned circuit (or to a load of negligible internal impedance).

which is independent of the rectification efficiency, and hence of the impressed signal, as it should be. (33) was, of course, derived under the assumption that the impressed voltage was not small, but it is not entirely accurate even when x is large, and its range of application is limited.

The initial, or small-signal, input resistance given by (34) can be expressed in another form as follows:

$$R_i' = \frac{R_1}{b} = \frac{R_1}{kR_1A_0'} = \frac{1}{kA_0'} \cdot \qquad (35)$$

The initial resistance is therefore inversely proportional to the parameter k and to the initial current through the diode, and depends only upon these two quantities.

Fig. 7 shows plots of (32) for $b=1$, 10, and 100, and we see a vast difference in the way in which the input resistance may vary with impressed voltage. If b is large, the resistance will rise as the signal increases, while if b is small the reverse change will occur. It would

appear from this figure that there must be a value of b which will make the input resistance almost independent of the impressed signal amplitude.

It has been pointed out that a large value of b improves the efficiency of rectification and that it also tends to make this efficiency nearly independent of the magnitude of b. However, if, for a given R_1, b is increased by using a positive bias, the initial current will go up and the initial resistance will go down. On the other hand, if R_1 and the positive bias are increased together, so as to keep A_0' constant, the initial resistance will be unchanged. From the standpoint of rectification efficiency, a value of $b=20$ is probably adequate, and if $R_1=50$ meghoms and $k=11.6$, the corresponding value of A_0' is 0.035 microampere and the initial input resistance is 2.5 megohms, assuming that the exponential law holds at very low currents. As the signal level is raised, the input resistance will also increase and will approach a limiting value of 25 megohms. This, of course, does not take into account the shunting effect of R_1, which is in parallel with the resistance of the diode. If higher values of initial resistance are required, b may be reduced, which will result in some decrease in sensitivity, especially at low signal voltages.

Permissible Internal Resistance in the Circuit Under Measurement

If a diode voltmeter is used to measure the voltage across a circuit having resistance for all frequencies, a serious error in the readings may easily occur, because the diode current is relatively quite large when the impressed voltage is at its positive peak. As has been pointed out in connection with (27) to (29), the amplitude of many of the harmonics may be almost as large as the fundamental, which, in turn, can be nearly twice the magnitude of the direct current. When $\cos \omega t=1.0$, all of the alternating-current components are in phase and a large current peak results.

In order to examine this case, it will be sufficient to consider a voltage E cost ωt applied through a resistance R_0 to the voltmeter terminals. While a full solution has not been obtained in explicit form, it is possible to set up equations which will give the value of R_0 required to reduce the rectified voltage ΔE_1 by 1 per cent. These equations will define the maximum safe circuit resistance, as seen from the voltmeter terminals, that can be used without appreciable error.

In this section, we shall let A_0 be the total direct current through R_1 which occurs when the voltage $E \cos \omega t$ is applied directly to the terminals 1–2 of Fig. 8, while a_0 is the change in this current which

occurs when the resistance R_0 is inserted. (a_0 will, of course, be negative.) Let

$$B_0 = A_0 + a_0 = \Delta A_0 \left(1 + N + \frac{b}{\beta x}\right), \tag{36}$$

in which

$$N = a_0/\Delta A_0. \tag{37}$$

For future convenience, we may note that

$$kR_1 a_0 = kR_1 \Delta A_0 \frac{B_0}{\Delta A_0} = \beta x N. \tag{38}$$

Also, let

$$R = \frac{R_1 R_0}{R_1 + R_0}. \tag{39}$$

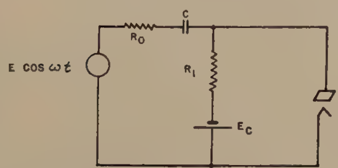

Fig. 8—The diode voltmeter connected to a generator with an aperiodic internal resistance R_0.

If a small amount of resistance is introduced at R_0, there will result, corresponding to (5), the following equation:

$$B_0 + s = G\epsilon^{-kE_c}\epsilon^{kE\cos\omega t}\epsilon^{-kR_1B_0}\epsilon^{-kRs}, \tag{40}$$

in which

$$s = A_1 \cos \omega t + A_2 \cos \omega t + \cdots = A_m \sum_{m=1}^{\infty} \cos m\omega t. \tag{41}$$

In view of (7) and (23), (40) can be written

$$(B_0+s)\epsilon^{kR_1a_0} = G\epsilon^{-kE_c}\epsilon^{-kR_1A_0}I_0(x)\left[1+2\sum_{n=1}^{\infty} f_n(x)\cos n\omega t\right]\epsilon^{-kRs}. \tag{42}$$

If the inserted value of R_0 is small, kRs will also be small, and the last factor in (42) can be replaced by $(1-kRs)$. Then, in view of (38) and (14), we may write

$$\frac{(B_0 + s)}{A_0}\epsilon^{\beta x N} = \left[1 + 2\sum_{n=1}^{\infty} f_n(x)\cos n\omega t\right][1 - kRs]. \tag{43}$$

The direct-current component of the right-hand side of this equation can be calculated if we assume that the amplitudes A_1, A_2, etc., are not appreciably affected by the small resistance introduced at R_0, and are therefore given by (29). Under this assumption, we can write for s

$$s = 2B_0 \sum_{m=1}^{\infty} f_m(x) \cos m\omega t, \tag{44}$$

and, substituting this into (43), it is evident that the direct-current component of the right side is

$$1 - 2kRB_0 \sum_{m=1}^{\infty} f_m^2(x) = 1 - \frac{2kRB_0}{I_0^2(x)} \sum_{m=1}^{\infty} I_m^2(x). \tag{45}$$

If x is large, I_m will also be large for a considerable range of m, and the computation of the summation indicated on the right of (45) would be somewhat tedious. We therefore make use of a theorem given on page 30 of Watson,[5] which states:

$$2 \sum_{m=1}^{\infty} (-1)^m J_m^2(z) = J_0(2z) - J_0^2(z). \tag{46}$$

Let $z = jx$, and substitute into (46), getting

$$2 \sum_{m=1}^{\infty} (-1)^m J_0^2(jx) = J_0(2jx) - J_0^2(jx), \tag{47}$$

and, in view of (9), this can be written

$$2 \sum_{m=1}^{\infty} I_m^2(x) = I_0(2x) - I_0^2(x). \tag{48}$$

Now, let

$$F(x) = \frac{I_0(2x)}{I_0^2(x)} - 1. \tag{49}$$

Then (45) reduces to

$$1 - kRB_0F(x) = 1 - \frac{R}{R_1} (\beta x + \beta x N + b)F(x), \tag{50}$$

which is the direct-current component of the right-hand side of (43).

Turning now to the left-hand side of (43), the direct-current component is

$$\frac{B_0}{A_0} \epsilon^{\beta x N} = \left(\frac{\Delta A_0 + a_0 + A_0'}{\Delta A_0 + A_0'}\right) \epsilon^{\beta x N} = \left(\frac{\beta x + \beta x N + b}{\beta x + b}\right) \epsilon^{\beta x N}. \tag{51}$$

Equating (50) and (51) and solving for R/R_1, there results

$$\frac{R}{R_1} = \frac{1}{F(x)(\beta x + b)} \left[\frac{1}{1 + \dfrac{\beta x N}{\beta x + b}} - \epsilon^{\beta x N} \right]. \tag{52}$$

In order that the approximations which have been made may be valid, it is necessary to consider only quite small values of N, such as -0.01. When this is done, (52) may be written

$$\frac{R}{R_1} = \frac{1}{F(x)(\beta x + b)} \left[1 + \frac{0.01\beta x}{\beta x + b} - \epsilon^{-\beta x/100} \right], \tag{53}$$

and we can still obtain an entirely sufficient degree of accuracy by dropping the small term in the center of the square brackets, and writing

$$\frac{R}{R_1} = \frac{1}{F(x)(\beta x + b)} \left[1 - \epsilon^{-\beta x/100} \right]. \tag{54}$$

β can be obtained from Fig. 4, and F(x) is given by Table I.

TABLE I

x	$F(x)$	x	$F(x)$
0	0	10.0	4.48
0.5	0.121	15.0	5.76
0.7	0.226	20.0	6.93
1.0	0.422	30.0	8.72
2.0	1.172	50.0	11.54
3.0	1.825	60.0	12.73
5.0	2.813	80.0	14.86
7.5	3.73	100.0	16.72

Plots of (54) for $b=1$ and $b=10$ are shown in Fig. 9. The points in the right-hand part of the drawing are experimentally determined values of R_0 for $b=10$. A heater voltage of 3.5 was used, which made $k=1.2$ In view of the small magnitude of the voltage increment being measured (one per cent of ΔE_1), and of the sensitiveness of the measurement to small values of harmonic content, these checks are felt to be satisfactory, especially in view of the fact that we are interested only in an approximate value of the quantity R/R_1. The circles indicate measurements made with filtered 60-cycle power supply, which gave trouble from voltage fluctuation, and the squares indicate measurements taken with a somewhat less-filtered battery-operated oscillator supply.

Very much larger values of R/R_1 are obtainable with $b=1$, provided the impressed alternating voltage is small. However, if x is greater than about 30, there is little difference between the curves for

the two values of b, and with $k=12$ this corresponds to a peak voltage of only 2.5 volts. Consequently, by using a negative bias to reduce b, we can considerably increase the value of internal circuit impedance that may be allowed for measurements of quite small voltages, but the

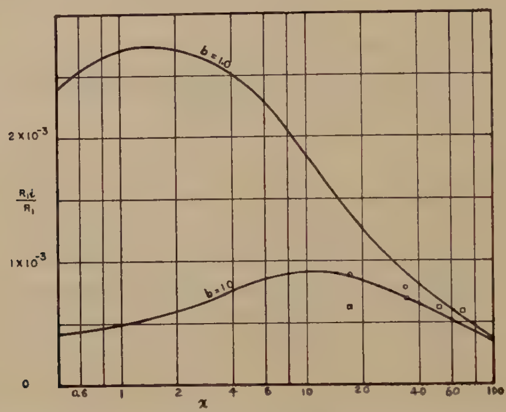

Fig. 9—Ratio of generator resistance to grid-leak resistance required to cause a reduction of 1 per cent in the efficiency of rectification.

advantage is entirely lost if the voltage is made large. This is not surprising, since it requires a change in bias of only -0.942 volt to reduce b from 10 to unity, and when the impressed voltage peak is several times as large as this we cannot expect the small alteration of the bias to have any great effect.

Proceedings of the Institute of Radio Engineers
Volume 26, Number 7 *July, 1938*

A METHOD OF DESIGNING SIMULATIVE NETWORKS*

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Summary—*In communication work it is often necessary to design a simple lumped network, the impedance of which approximates at all frequencies that of a certain distributed circuit, such as a transmission line. The present treatment includes a clarified method of making such a design.*

Former design methods have the common fault that negative circuit constants may be called for. This paper presents a simple method by which the difficulty is avoided. Moreover the best design under the particular circumstances in question may be arrived at quickly and directly.

The impedance functions of two common forms of simulative networks are developed and design equations for the establishment of the parameters are presented. Since four elements are to be established it is possible to specify four arbitrary conditions. This is most conveniently handled by specification of the vectorial impedance at two frequencies. The major portion of this paper is concerned with the relation which must exist between these impedances in order that the calculated circuit parameters will all be positive. For the sake of completeness conditions leading to negative parameters are also considered.

Introduction

A METHOD of designing networks which serve as wave-filter-terminal transducers, wave-filter-impedance simulators, and line-impedance simulators has been presented by Zobel.[1] In the present paper the method as applied only to the simulation of lines will be considered. This method is applicable to any problem in which a given impedance curve of appropriate form is to be approximated.

Zobel further simplifies the problem by dividing it into two parts. First, a basic network is designed from considerations of filter theory which approximates the desired impedance at all except low frequencies. Second, a supplementary network or excess simulator is designed to correct the deviation in impedance at the lower frequencies. Since in nonloaded lines the basic network is simply a pure resistance and since the design of the basic network for loaded lines is reasonably easy, the entire problem hinges on the design of the excess simulator. The excess simulator usually takes the form of Fig. 1(a), sometimes that of 1(b). Each network consists of four elements, but the two give somewhat different impedance characteristics.

* Decimal classification: R140. Original manuscript received by the Institute, August 5, 1937; revised manuscript received by the Institute, January 13, 1938.

[1] "Extensions to the theory and design of electric wave filters," *Bell Sys. Tech. Jour.*, vol. 10, pp. 284–341, April, (1931).

Zobel presents the circuit of Fig. 1(a) and a method of determining the values of the four elements. The required data are the values of the vectorial impedance at two frequencies which the excess simulator is required to give. The circuit of Fig. 1(b) is useful in some cases when that of Fig. 1(a) is inapplicable.

The solution of the equations presented by Zobel is perfectly straightforward but unfortunately it often happens that one or more of the four circuit elements required has a negative value. This of

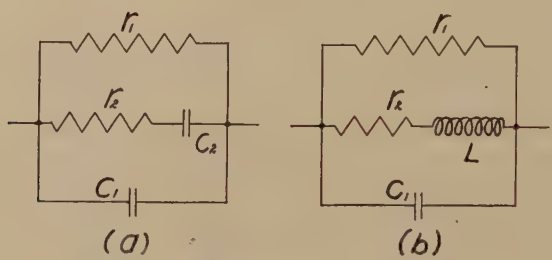

Fig. 1—Excess-simulator networks.

course is not physically possible and no remedy for the situation is obvious. For, while it is usually permissible to make small alterations in the impedances substituted in the equations, it is not apparent what alteration will yield the desired positive elements.

A method of dealing with this situation will now be developed which predicts from the given data with little effort which, if any, of the elements is required to be negative and what modification will yield the most satisfactory set of positive elements. The derivation is carried out in terms of admittance because the mathematics is considerably simplified thereby.

Design Procedure

Ordinarily a curve such as that of Fig. 2 is at hand representing the impedance characteristic which the network is to approximate. These data are easily transformed into the polar plot of Fig. 3 and thence to the admittance plot of Fig. 4.

A moment's consideration shows that either circuit of Fig. 1 will, at least for some values of the parameters, approximate the curves shown. In all cases the impedance is zero at infinite frequency, finite and pure resistive at zero frequency, and capacitative at finite frequencies. The circuit of Fig. 1(b) gives a loop in its polar impedance plot for low frequencies unless the inductance is relatively small. Accordingly the circuit of Fig. 1(a) is usually to be preferred.

Since the impedance of the networks equals that of the given empirical data at infinite frequency and can be made equal to it at two finite frequencies a reasonably good fit at all frequencies can be expected. Actual test justifies this expectation.

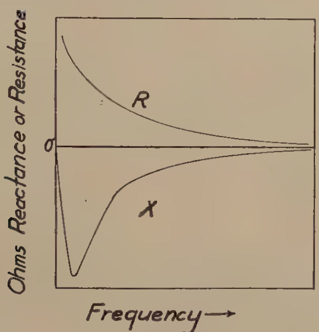

Fig. 2—Given impedance characteristic to be approximated.

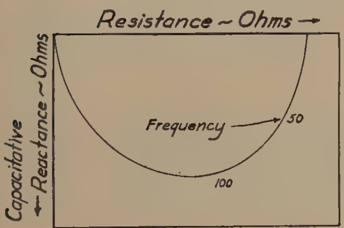

Fig. 3—Polar impedance transformation.

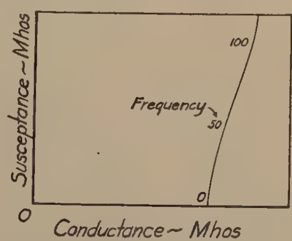

Fig. 4—Admittance transformation.

The admittance of Fig. 1(a) is given by

$$Y = j\omega c_1 + \frac{1}{r_1} + \frac{1}{r_2 + \dfrac{1}{j\omega c_2}} = j\omega c_1 + \frac{1}{r_1} + \frac{j\omega c_2}{1 + j\omega c_2 r_2}$$

$$Y = \frac{\dfrac{1}{r_1} + \dfrac{j\omega}{r_1}(c_1 r_1 + c_2 r_2 + c_2 r_1) - \omega^2 c_1 c_2 r_2}{1 + j\omega c_2 r_2}.$$

Let[2]

$$Y = \frac{a_0 + jna_1 - n^2 a_2}{1 + jnb_1}.$$

[2] This substitution is similar to one made by Zobel.

The two admittance forms just given are identical if $\omega = n\omega_0$

$$\text{and} \left\{ \begin{aligned} a_0 &= \frac{1}{r_1} \\ na_1 &= \frac{n\omega_0}{r_1}(c_1r_1 + c_2r_2 + c_2r_1) \\ n^2a_2 &= n^2\omega_0{}^2c_1c_2r_2 \\ nb_1 &= n\omega_0c_2r_2 \end{aligned} \right\} \text{ or } \left\{ \begin{aligned} r_1 &= \frac{1}{a_0} \\ c_1 &= \frac{a_2}{\omega_0b_1} \\ r_2 &= \frac{b_1{}^2}{a_1b_1 - a_0b_1{}^2 - a_2} \\ c_2 &= \frac{a_1b_1 - a_0b_1{}^2 - a_2}{\omega_0b_1} \end{aligned} \right\}.$$

Suppose that the empirical curve in the form of Fig. 4 shows an admittance Y_0 for $\omega = \omega_0$ and Y_1 for $\omega = n\omega_0$. These conditions are sufficient to determine the network parameters r_1, r_2, c_1, and c_2. Thus we have

$$Y_0 \equiv G_0 + jB_0 = \frac{a_0 + ja_1 - a_2}{1 + jb_1}$$

and

$$Y_1 \equiv G_1 + jB_1 = \frac{a_0 + jna_1 - n^2a_2}{1 + jnb_1}.$$

Expanding

$$\left\{ \begin{aligned} G_0 + jB_0 + jb_1G_0 - b_1B_0 &= a_0 + ja_1 - a_2 \\ G_1 + jB_1 + jnb_1G_1 - nb_1B_1 &= a_0 + jna_1 - n^2a_2 \end{aligned} \right\}.$$

Simplifying

$$\begin{aligned} a_0 + 0 - a_2 + B_0b_1 &= G_0 \\ 0 + a_1 + 0 - G_0b_1 &= B_0 \\ a_0 + 0 - n^2a_2 + nB_1b_1 &= G_1 \\ 0 + na_1 + 0 - nG_1b_1 &= B_1. \end{aligned}$$

The above is seen to constitute a system of four simultaneous linear equations in four unknowns. That is, a_0, a_1, a_2, and b_1 are expressed in terms of G_0, G_1, B_0, B_1, and n. The solution is most conveniently carried out by the familiar determinant method.

Symbolically we may write

$$a_0 = \frac{N_1}{\Delta}; \qquad a_1 = \frac{N_2}{\Delta}; \qquad a_2 = \frac{N_3}{\Delta}; \qquad b_1 = \frac{N_4}{\Delta}$$

the area bounded by certain of these curves the sign, positive or negative, of all four circuit parameters is immediately fixed.

The boundary curves are very simply determined by equating the various determinant terms to zero. Thus as $\Delta \to 0$, then $r_1 \to 0$, $r_2 \to 0$, and $c_2 \to \infty^2$. Accordingly values of Y_1 for which $\Delta = 0$ are on a boundary on one side of which r_1 and r_2 are positive, on the other side of which r_1 and r_2 are negative. c_2 does not change sign here because of the square term. When $N_1 \to 0$; $r_1 \to \infty$ and a reversal in the sign of r_1 is noted. If $N_3 \to 0$; $c_1 \to 0$ indicating reversal of the sign of c_1. For $N_4 \to 0$; $c_1 \to \infty$, $c_2 \to \infty$, and $r_2 \to 0$ as the square. Reversal in the sign of c_1 and c_2 but not of r_2 is indicated.

From the above it is seen that the boundary curve for the reversal of r_1 and r_2 has the equation $(n^2-1)n(G_1-G_0)=0$. Since $0 \neq n \neq 1$ this reduces to the simple form $G_1 = G_0$.

For the reversal of r_1 alone we have

$$(B_1 - nB_0)^2 + (G_1 - G_0)(G_1 - n^2G_0) = 0$$

or

$$(B_1 - nB_0)^2 + \left(G_1 - \frac{1+n^2}{2} G_0\right)^2 = \left(\frac{1-n^2}{2} G_0\right)^2,$$

obviously the equation of a circle.

For reversal of c_1 and c_2 the equation is $(1-n^2)(B_1-nB_0)=0$ or simply $B_1 = nB_0$. The remaining boundary is that for which only c_1 changes sign. The equation is

$$n(G_0 - G_1)^2 = (nB_0 - B_1)(nB_1 - B_0)$$

or

$$(G_1 - G_0)^2 + \left(B_1 - \frac{n^2+1}{2n} B_0\right)^2 = \left(\frac{n^2-1}{2n} B_0\right)^2$$

again a circle. These equations are shown plotted in Fig. 5.

Evidently if Y_1 lies within one of the areas of Fig. 5 the four parameters r_1, r_2, c_1, and c_2 are all positive. When the value of Y_1 is shifted the parameters change sign as the various boundaries are crossed. These areas are identified, without presentation of the argument required. The points which appear in this diagram are the results of actual computation for the circuit values indicated. Naturally they check the identification given. It is to be noted that a crosshatch line of a given slope indicates that a certain element is negative in a region.

It will be observed that nine areas in all are formed. In four areas two elements are negative, in another four one element is negative, and in one area no element is negative. Although the figures are drawn

for $n=3$ they are similar for other values of n. The construction is valid for n less than one but the appearance is quite different.

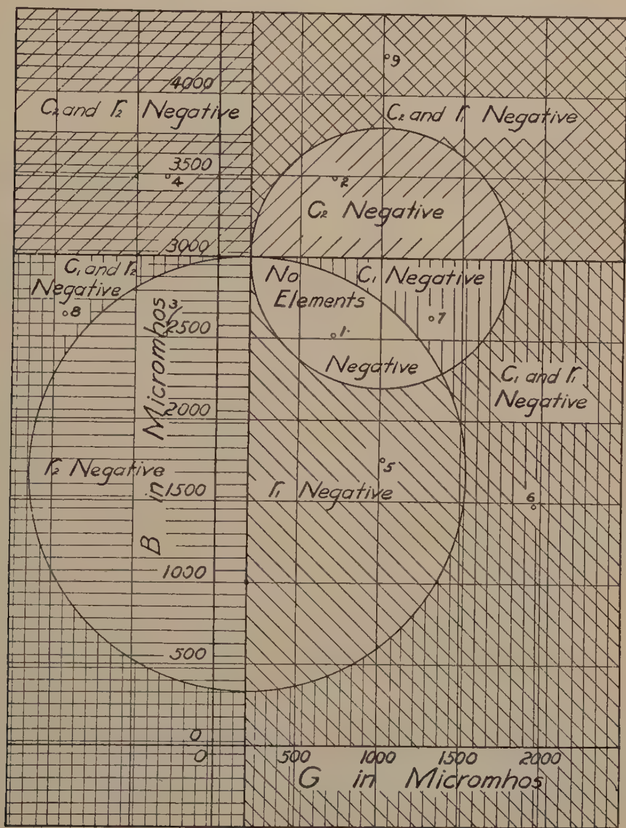

Fig. 5—Identification of areas.

Point No.	c_1 μf	c_2 μf	r_1 ohms	r_2 ohms	Y at 100~	Y at 300~
1	0.985	0.666	12500	750	$200+j1000$	$707+j2520$
2	2.20	−0.666	12500	750	$200+j1000$	$707+j3480$
3	0.985	0.666	3122	−750	$200+j1000$	$-307+j2520$
4	2.20	−0.666	3122	−750	$200+j1000$	$-307+j3480$
5	0.526	1.333	−7350	600	$200+j1000$	$1022+j1758$
6	−0.533	2.300	−5410	200	$200+j1000$	$1955+j1473$
7	−0.680	2.300	21750	75	$200+j1000$	$1316+j2637$
8	−0.680	2.300	2822	−75	$200+j1000$	$-916+j2637$
9	2.660	−1.333	−7350	600	$200+j1000$	$1022+j4242$

The application of the method is quite simple. G_0, B_0, G_1, B_1 and n are known from the empirical data such as that of Fig. 4. From only G_0, B_0, and n the complete diagram is easily constructed without

The equation above is rather difficult to compute. By considera-tion of its special properties, however, the labor can be reduced to a

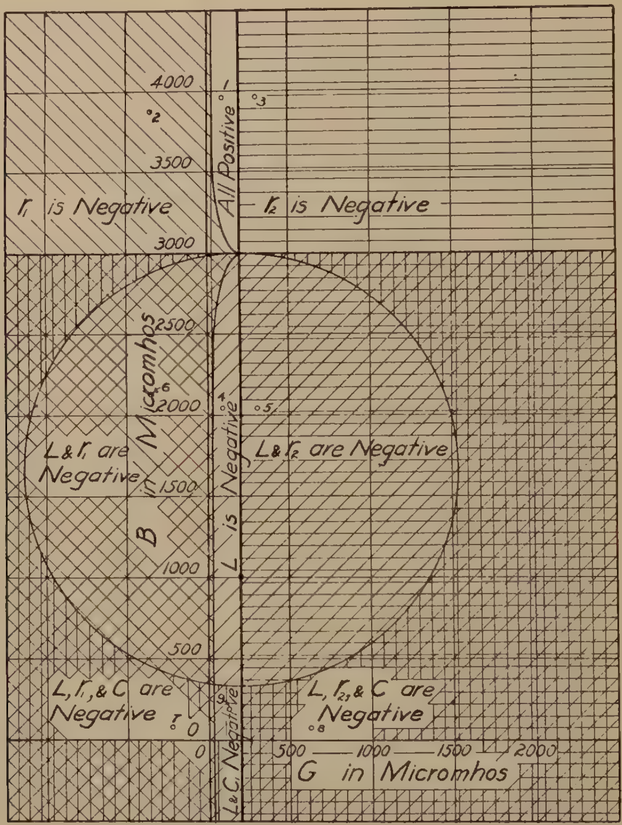

Fig. 6—Identifications of areas for alternative network.

Point No.	c_1 μf	L	r_1	r_2	Y at 100~	Y at 300~
1	2.17	4.0	12340	750	$200+j1000$	$94+j3964$
2	2.32	0.75	−1468	900	$200+j1000$	$-353+j3870$
3	2.17	4.0	12340	−750	$200+j1000$	$306+j3964$
4	2.17	−4.0	12340	750	$200+j1000$	$94+j2036$
5	2.17	−4.0	12340	−750	$200+j1000$	$306+j2036$
6	0.865	−0.75	−1468	900	$200+j1000$	$-353+j2130$
7	−0.186	−0.60	−3470	330	$200+j1000$	$-225+j78$
8	−0.186	−0.60	1450	−330	$200+j1000$	$625+j78$
9	0.089	−0.75	8640	75	$200+j1000$	$125+j185$

reasonable amount. B_1 is real only for values of G_1 between G_0/n^2 and G_0, and the curve is assymptotic to the line $G_1=G_0/n^2$. Moreover, the curve is symmetric about $B_1=nB_0$ and has a cusp at the point

(G_0, nB_0). This and other boundary curves are plotted in Fig. 6. It is worth noting that a very rough sketch of the cubic curve is often sufficient for the purpose.

In the same figure the various areas formed are identified and actual computed points are presented. Again a crosshatch line of a given slope indicates the fact that a certain element is negative in a region. The diagram is strikingly different from that for the other circuit. In particular two areas occur for which three elements are negative, and correspondingly only three areas occur for one and for two elements negative. The shape of the all-positive area is particularly interesting.

Limitations of the Method

It will be observed that the present treatment is confined to two specific four-element networks and, through degenerate cases, three three-element nets. No mention has been made of any other.

It can be shown that there are just 42 irreducible four-element networks of resistance, inductance, and capacitance. Of these only two are free from resistance. Unfortunately no method applying to all in general has been found. Any individual circuit can be treated by a method analogous to that presented here, though in some cases the algebra is more involved.

It will be observed that for a given Y_0 and n a two-element network gives only one point for Y_1. A three-element network gives a locus which is a portion of a curved or straight line, while a four-element net gives a definite area. Immediately the question arises as to the behavior of a five-element net. It can be shown that one additional fact is necessary to obtain a solution. Thus if the direct-current resistance and the admittance at one frequency are known for a five-element network a definite area for Y_1 will be obtained. For a six-element network two additional facts are necessary, and it is often difficult to establish these facts so that the possibility of negative elements is precluded. Certainly the procedure is not simple.

Appendix

An example demonstrating the method

To clarify the method and to indicate its application and limitations an actual problem will be solved. The data shown in Fig. 7 were taken on an artificial loaded cable of 24 sections each representing 6000 feet of 19-gauge cable loaded with 88-millihenry coils at that spacing and terminated midsection. The far end of the cable was terminated in a basic network calculated from the constants of the line.

400-cycle admittance to the boundary of the positive area. A decrease in the value of G as well as in B seems desirable.

The equation of the boundary in question is

$$(G_1 - G_0)^2 + \left(B_1 - \frac{n^2+1}{2n} B_0\right)^2 = \left(\frac{n^2-1}{2n} B_0\right)^2.$$

If, somewhat arbitrarily, we choose 490 as the value for G_1 we obtain the value 3582 for B_1. We are now in a position to compute the size of the required elements. It will be recalled that

$$r_1 = \frac{\Delta}{N_1}, \qquad c_1 = \frac{N_3}{\omega_0 N_4}, \qquad r_2 = \frac{\Delta N_4^2}{\Delta N_2 N_4 - N_1 N_4^2 - \Delta^2 N_3}, \quad \text{and}$$

$$c_2 = \frac{\Delta N_2 N_4 - N_1 N_4^2 - \Delta^2 N_3}{\omega_0 \Delta^2 N_4}.$$

Moreover the boundary in question requires that $C_1 = 0$. Accordingly $N_3 = 0$ and only four of the five determinant terms need be computed. They are

$$\Delta = (n^2-1)n(G_1 - G_0) = 10.11 \times 3.33 \times \frac{490-250}{10^6} = 8090. \times 10^{-6}$$

$$N_1 = -n\{(B_1 - nB_0)^2 + (G_0 - G_1)(n^2 G_0 - G_1)\} = 1.825 \times 10^{-6}$$

$$N_2 = (n^2-1)(nG_1B_0 - G_0B_1) = 8.775 \times 10^{-6}$$

$$N_4 = (n^2-1)(nB_0 - B_1) = 180. \times 10^{-6}.$$

The element sizes follow immediately. Several of the equations are simplified by the fact that $N_3 = 0$. We have

$$r_1 = \frac{\Delta}{N_1} = 4440 \text{ ohms}$$

$$r_2 = \frac{\Delta N_4}{\Delta N_2 - N_1 N_4} = 20.8 \text{ ohms}$$

$$c_2 = \frac{\Delta N_2 - N_1 N_4}{\omega_0 \Delta^2} = 1.43 \text{ microfarads.}$$

The problem is now formally solved. For the sake of a check, however, it is well to compute the performance of the final design. The computation of several typical points and a convenient computing form are presented below. The over-all performance is indicated in Fig. 7.

The impedance and admittance of the series combination of r_2 and c_2 are first computed. The conductance of r_1 is then added to obtain the total admittance. Finally the over-all impedance is obtained by another reciprocal transformation.

Frequency	Z_1 ohms	$\|Z_1\|$	Y_1 micromhos	Y	$\|Y\|$	Z ohms
100	$20.8-j1110$	1110	$16.9+j900$	$242+j900$	931	$279-j1038$
120	$20.8-j925$	925	$24.3+j1080$	$249+j1080$	1109	$202-j879$
200	$20.8-j555$	555	$67.1+j1795$	$292+j1795$	1820	$88-j543$
400	$20.8-j278$	279	$268+j3585$	$493+j3585$	3622	$37-j273$
1000	$20.8-j111$	113	$1629+j8690$	$1854+j8690$	8890	$23-j110$

It is to be noted that the impedance approaches the parallel value of the two resistors instead of zero as the frequency increases. Such is always the case when C_1 is zero. In the present case such behavior seems desirable.

7:30 P.M., E.S.T., in this case, the residual signal strength at midnight is considerably greater than it was at 6 A.M.

The marked increase in field strength beginning at about 12:30 A.M., February 21, is an example of a phenomenon to be discussed later. Unfortunately the transmitter was shut down at 1:00 A.M., before the signal strength had returned to normal.

An increase in terrestrial magnetic activity has two main effects. Critical frequencies are lowered so that the reflected signal appears later in the morning and disappears earlier in the evening, as shown by Kirby, Gilliland, Judson, and Smith.[1] On a day of magnetic charac-

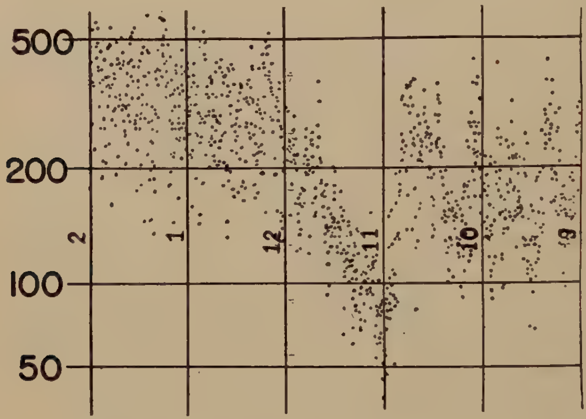

Fig. 2—Dellinger phenomenon.

ter 1 ("moderately active") in February and March, the signal has been reflected for about two hours less than on a 0 day. A magnetic storm frequently keeps the critical frequency below 9.57 megacycles, so that no reflection is observed for a day or two at a time. Also, in general, the variation in signal strength is more marked. Even a moderate magnetic day of, say, 100-γ magnitude, can easily double the spread between maximum and minimum field strengths recorded.

Dellinger fade-outs are observed occasionally, in less complete form than when observed over long distances. Presumably these fade-outs are caused by a type of solar radiation of very sudden onset which expends its energy in ionizing a fairly thin layer at a relatively low altitude where collisional friction is important. The vertically reflected signal is absorbed less than any other because it traverses the absorbing stratum in the most direct path. These fade-outs are characterized

[1] S. S. Kirby, T. R. Gilliland, E. B. Judson, and N. Smith, *Phys. Rev.*, vol. 48, p. 849. (1935); vol. 50, p. 258, (1936).

by the sudden onset and gradual decrease of absorption shown in the example in Fig. 2.

On two occasions fade-outs of a different sort have been observed which are definitely local in nature. Fig. 3 shows a fade-out which was of small magnitude and duration in Cambridge. This was observed as a "deep fade on W1XK but not on W8XAL" by Kirby and Smith, who have been recording the signal strengths of these stations for nearly three years. This fade-out must be ascribed to an ionized absorbing region of small area located over W1XK, possibly on the Washington side. The example in Fig. 4 was not noted in Washington,

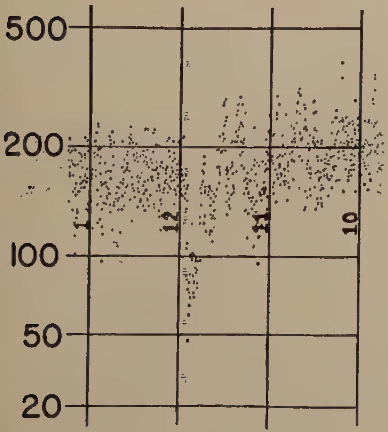

Fig. 3—Local fade-out.

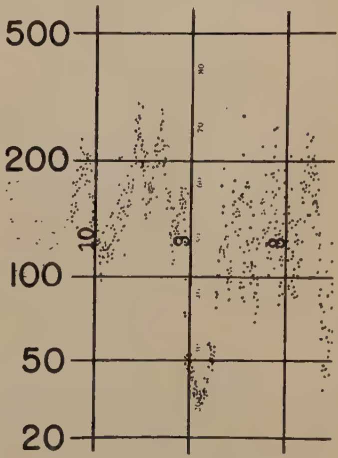

Fig. 4—Local fade-out.

although much more marked in Cambridge. It was probably due to an ionized region so located as to attenuate both ascending and descending rays.

Apparently related to these local fade-outs are bursts of signal received in the evening or early morning when F_2 reflection is absent. Signals of this type are well above the normal level of the ground wave and scattered rays, and normally exhibit the same symmetry as the daytime fade-outs. Most cases last for about the same length of time, from five to fifteen minutes. An example of one of these bursts appears at the left of Fig. 1, and a smaller and more typical one is shown in Fig. 5.

These bursts are probably produced by the same mechanism as that causing the fade-outs. The absorption of a wave of fixed frequency increases with the density of ionization, becoming very great (if in a low layer) when the ionization is nearly dense enough to reflect the

the meteor, along which the ionization is intense. In all directions perpendicular to this path the density of ionization decreases with distance from the axis. A ray approaching this ionized region in a plane containing the axis will be refracted so that the direction of propagation becomes parallel to the axis, and then will escape from the ionized region along a similar path. This is exactly equivalent to specular reflection from a cylinder whose radius is somewhat smaller than the nearest approach of any ray to the axis.

Under these conditions it is seen that no energy will be reflected to its source unless a perpendicular may be passed from the source to the path of the meteor; and that if this condition is fulfilled the energy so reflected may be calculated very easily.

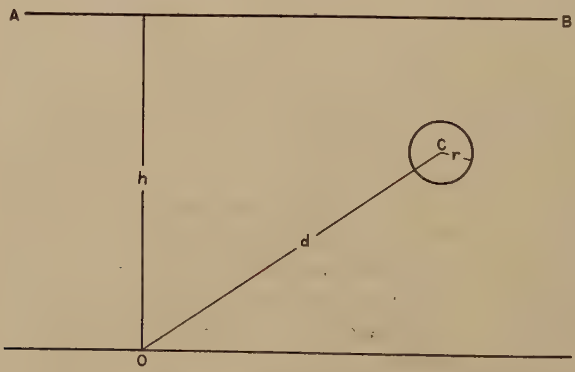

Fig. 6—Notation used for calculating relative reflected energy.

Consider Fig. 6 where AB represents the equivalent reflecting plane for an ionized layer and the circle with center at C represents the equivalent reflecting cylinder containing a meteor track along its axis. The energy radiated from 0 and reflected from the plane AB, to fall upon a small area also at 0 is,

$$E_p = \frac{K_1}{h^2}. \tag{1}$$

For energy reflected to 0 from the cylinder C the density in the plane of the paper is nearly proportional to r/d and the energy distribution in the plane perpendicular to the paper is the same as for the case of a plane surface, so that

$$E_c = \frac{K_1}{d^2}\cdot\frac{r}{d} = \frac{K_1 r}{d^3}. \tag{2}$$

The ratio of the energy reflected by C to that reflected by AB is

$$R = \frac{rh^2}{d^3}. \tag{3}$$

In a typical case we may compare the energy reflected from a cylinder at a distance of 100 kilometers with a reflection from the F region at, say, 250 kilometers. Let

$$d = 100 \text{ kilometers}$$
$$h = 250 \text{ kilometers}$$
$$r = 0.25 \text{ kilometer.}$$

Then

$$R = \frac{0.25 \cdot 6.25 \cdot 10^4}{10^6} = 0.0156.$$

The square root of this value is the ratio of field strengths, 12.5/100. This compares favorably with the experimental values of the field strength in our "bursts."

From the appearance of the original records it seems that a signal having only two or three per cent of the field strength of that reflected from the F layer might easily be lost in the background due to scattering and ground waves. Since R in (3) varies inversely as d^3, we should not expect to notice the reflection due to the passage of a meteor much farther than 200 or 250 kilometers from the observer.

It is now necessary to determine the size of a meteor which might ionize such volumes to the degree necessary to reflect a 10-megacycle wave. To ionize a region 0.5 kilometer in radius and 100 kilometers long,

$$m = \frac{2\,K.E.}{V^2}$$

$$= \frac{2 \cdot 1.6 \cdot 10^{-12} \dfrac{\text{ergs}}{\text{electron volt}} \cdot 15 \dfrac{\text{electron volts}}{\text{ion}} \cdot 0.8 \cdot 10^{17}\ \text{cm}^3 \cdot 10^6 \dfrac{\text{ions}}{\text{cm}^3}}{1.6 \cdot 10^{13} \dfrac{\text{cm}^2}{\text{sec}^2}}$$

$$= 250 \text{ milligrams} \tag{4}$$

assuming

1. the first ionization potential of the average atom in the atmosphere to be 15 electron volts,
2. the velocity of the meteor to be 40 kilometers per second,

Let us make the conservative assumption that the rate of producing ions in the E region is that which we would have if the energy of the meteor were expended uniformly throughout a vertical distance (not path length) of 100 kilometers. Then,

$$\frac{0.93 \cdot 10^6 \text{ ions/cm}^2 \text{ sec}}{10^7 \text{ cm}} = 0.093 \text{ ions/cm}^3 \text{ sec.} \tag{10}$$

The energy producing these new ions is derived from the advent of some 10^{11} meteors per day, practically all of which are of telescopic magnitude. This corresponds to about 100 meteors per second within the region from which we may expect to receive a reflection. Thus, on the average, a modest rate of diffusion is sufficient to account for a nearly uniform distribution of this new ionization.

The density of ionization N, which can be maintained by N_1 new ions per second is,

$$N = \sqrt{\frac{N_1}{\alpha}} \tag{11}$$

where α is the recombination coefficient. Skellett[2] shows that α appears to be smaller than 0.2×10^{-8} in the E region, while Lenz[8] deduces a probable value of approximately 10^{-10} at a height of about 100 kilometers. If we adopt 10^{-9} as a conservatively high value,

$$N = \sqrt{\frac{9.3 \cdot 10^{-2}}{10^{-9}}} = 10^4 \text{ ions/cm}^3 \tag{12}$$

which corresponds to a value of the E-layer critical frequency of 0.9 megacycle. While the experimental evidence is not very complete, apparently the ionization in the E layer remains at about this level during the night; a level which could be maintained by the constant bombardment of meteors.

The need of observations of the direction of arrival of the bursts of reflection is obvious. These observations should be co-ordinated with pulse measurements of effective heights (or, in this case, effective distances) of reflection, and with visual observation of meteors. Plans for such a program are being formulated at this laboratory.

During the short period covered by the data on which this paper is based, inclement weather prevented any consistent attempt to observe meteors. One meteor, described as of about -1 magnitude, was sighted by a friend of the writer. The time when it was seen could not be determined more closely than within about fifteen minutes, but

[8] E. Lenz, *Zeit. für Phys.*, vol. 76, p. 660, (1932).

within that limit checks with the time of a "burst" of smaller amplitude than that shown in Fig. 5. The fall of a meteorite at a point about 25 kilometers north of the receiver was reported during the period in which these observations were made. No corresponding burst of signal was noted. This agrees with the deduction that proper orientation of the meteor's path is a condition necessary to this phenomenon. The path of this meteorite could not have fulfilled our requirement except near the surface of the earth, if at all. At such a low level the short mean free path of an electron, with the high absorption coefficient accompanying it, would permit the formation of a "cylinder" of ionization having only a small radius, and its collapse would be almost instantaneous; while the reflection of radio-frequency energy from this cylinder would be subject to enormous attenuation in the outer regions of small ionization density.

Acknowledgment

The writer is deeply grateful to Professor Harry Rowe Mimno and Professor F. V. Hunt of Cruft Laboratory, and to Dr. T. E. Sterne of the Harvard Observatory for their generous consideration and criticism of the ideas presented herein.

Appendix

Using the notation of Fig. 7, we wish to determine the number of meteors which pass through the region RT in such a way that a perpendicular may be dropped from O to the path of the meteor, with the intersection within the region.

This number may best be obtained in comparison with the number of meteors falling through the upper surface of the same region, regardless of their orientation. The cylindrical shape of this space is determined by considering that its vertical limits are the heights at which average meteors appear and disappear; while we are not interested in meteor tracks which are at too great a distance from O.

We must assume a perfectly random distribution of meteors; that is, a meteor is equally likely to come from any point in the celestial sphere and to pass through any point in the earth's atmosphere. Astronomical evidence indicates that this assumption is justified, for the number of meteors in the recognized showers is small in comparison with the total.

Let AB be one of a bundle of parallel rays, representing possible meteor paths, with a perpendicular from O intersecting it at B. Considering the family of possible paths parallel to AB, the intersections must all lie in a plane surface which makes an angle α with the vertical.

where $a=r/h$. This integral can be solved by putting $\tan^2\alpha=u$. Then

$$d\alpha = \frac{du}{2\sqrt{u}(1+u)}$$

and

$$\frac{M}{K_2dh} = \int_0^{a^2} \frac{a^2-u}{\sqrt{a^2-u}} \cdot \frac{du}{2(1+u)\sqrt{u}} \cdot \tag{17}$$

Dividing the numerator by $(1+u)$ we have

$$\frac{M}{K_2dh} = -\int_0^{a^2} \frac{du}{2\sqrt{u}\sqrt{a^2-u}} + \int_0^{a^2} \frac{(a^2+1)du}{2(1+u)\sqrt{u}\sqrt{a^2-u}} \cdot$$

In the first of these integrals we may substitute $u=v^2$; and in the second $1+u=1/z$. Then, as $du=2vdv$ in the first and $-dz/z^2$ in the second

$$\frac{M}{K_2dh} = -\int_0^a \frac{2vdv}{2v\sqrt{a^2-v^2}} - \frac{a^2+1}{2}\int_1^{1/(1+a^2)} \frac{-zdz}{z^2\sqrt{\dfrac{1-z}{z}}\sqrt{a^2-\dfrac{1-z}{z}}}$$

$$= -\int_0^a \frac{dv}{\sqrt{a^2-v^2}} - \frac{a^2+1}{2}\int_1^{1/(1+a^2)} \frac{dz}{\sqrt{1-z}\sqrt{a^2z+z-1}}$$

$$= -\int_0^a \frac{dv}{\sqrt{a^2-v^2}}$$

$$+\frac{a^2+1}{2}\int_{1/(1+a^2)}^{1} \frac{dz}{\sqrt{-z^2(a^2+1)+z(a^2+2)-1}} \cdot \tag{18}$$

Both of these forms are in the usual tables, so that we find

$$\frac{M}{K_2dh} = \left[\cos^{-1}\frac{v}{a}\right]_0^a$$

$$+\frac{a^2+1}{2}\left[\frac{1}{\sqrt{a^2+1}}\sin^{-1}\frac{2z(a^2+1)-(a^2+2)}{\sqrt{(a^2+2)^2-4(a^2+1)}}\right]_{1/(1+a^2)}^{1}$$

$$= \left[\cos^{-1}\frac{v}{a}\right]_0^a + \frac{\sqrt{a^2+1}}{2}\left[\sin^{-1}\frac{(a^2+1)z-(a^2/2+1)}{a^2/2}\right]_{1/(1+a^2)}^{1}$$

$$= -\frac{\pi}{2} + \frac{\sqrt{a^2+1}}{2}\left[\frac{\pi}{2}+\frac{\pi}{2}\right]$$

$$= \frac{\pi}{2}\left(\sqrt{a^2+1}-1\right). \tag{19}$$

Substituting r/h for a,

$$\frac{M}{K_2dh} = \frac{\pi}{2h}(\sqrt{r^2+h^2}-h)$$

or

$$M = \frac{\pi K_2 d}{2}(\sqrt{r^2+h^2}-h). \tag{20}$$

This number must be compared with the total number of meteors which fall through a circular area having the same radius as the region we have just considered.

The projection of this circle upon a plane perpendicular to the path of the meteor is an ellipse of major axis $2r$ and minor axis $2r \sin \alpha$.

The total number of meteors passing through this surface is, using the same terminology as in (14),

$$M_2 = \pi r^2 \sin \alpha \cdot 2\pi R M_0 \cos \alpha \, d\alpha \tag{21}$$

$$= \frac{\pi K_2}{2} r^2 \sin \alpha \cos \alpha \, d\alpha. \tag{22}$$

Integrating through all values of α

$$M_T = \frac{\pi K_2 r^2}{2} \int_0^{\pi/2} \sin \alpha \cos \alpha \, d\alpha \tag{23}$$

$$= \frac{\pi K_2 r^2}{2} \left[\frac{\sin^2 \alpha}{2} \right]_0^{\pi/2}$$

$$= \frac{\pi K_2 r^2}{4}. \tag{24}$$

Dividing (20) by this quantity

$$\frac{M}{M_T} = \frac{2\pi K_2 d(\sqrt{r^2+h^2}-h)}{\pi K_2 r^2}$$

$$= \frac{2d}{r^2}(\sqrt{r^2+h^2}-h). \tag{25}$$

The limits we must set for this region are somewhat uncertain, but we can establish them with sufficient accuracy. We may say that a meteor probably does not interest us if it is more than 200 kilometers away, and so set r as of the order of 150 kilometers. h and d may be taken from astronomical data, under the assumption that light will be

usable frequencies were considerably increased during occasional irregular periods by reflections from clouds of sporadic E layer. The E-layer graphs are included in Fig. 2 for convenience in determining multihop transmission conditions.

In Figs. 1 and 2 the curves representing F-layer and F_2-layer frequencies show further flattening out diurnally with blunt maxima

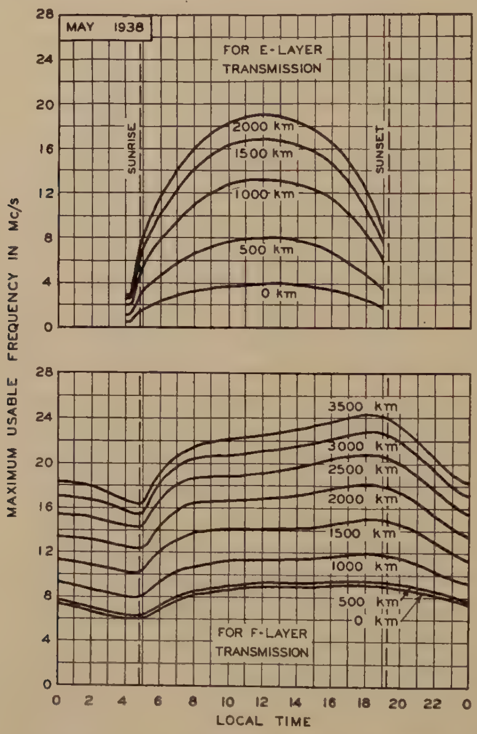

Fig. 2—Maximum usable frequencies for E- and F-layer transmission for the latitude of Washington; average for undisturbed days of May, 1938. The hour to be used is local time where the waves are reflected from the layer.

before sunset and minima as usual a half hour or more before sunrise. The F_1-layer critical frequencies were well defined during the daytime.

A severe ionosphere storm occurred on May 11 and 12. This storm was preceded by abnormally high f_{F_2} during the forenoon and by strong hissing static at high and broadcast frequencies about noon of May 11.

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ionosphere at Washington, D. C. For earlier publications on this subject see Proc. I.R.E., vol. 25, pp. 823–840; July, (1937), and a series of monthly reports beginning in Proc. I.R.E., vol. 25, pp. 1174–1191; September, (1937). Publication approved by the Director of the National Bureau of Standards of the U. S. Department of Commerce.

After this the f_{F_2} varied irregularly and rose to still higher values and h_{F_2} also rose abnormally. At about 1600 E.S.T. the normal stratification of the entire ionosphere almost completely disintegrated and a period of severe turbulence set in. During this period irregular, weak, jumpy reflections were observed at virtual heights varying erratically from 100 to 800 kilometers. During the late evening conditions became somewhat less erratic but reflections were complex, f_F abnormally low and h_F abnormally great. On May 12 the low f_{F_2}, great h_F, and high absorption continued all day without the marked turbulence. The

TABLE I

Ionosphere Storms (approximately in order of severity)

Date and hour E.S.T.	h_F before sunrise, km	Minimum $f_F{}^x$ during day (before sunrise), kc	Noon $f_{F_2}{}^x$ kc	Magnetic character[1]		Ionosphere character[2]
				00–12 G.M.T.	12–24 G.M.T.	
May 11 after 1200	—	—	10200	0.4	1.9	2
May 12	440	4400	6100*	1.7	1.0	2
May 13 until 0400	342	3900	—	0.4	0.0	0.5
May 14 after 0700	—	—	6450	0.8	1.2	1
May 15 until 1100	394	4100	—	1.0	0.5	0.5
May 4	362	5200	6950	0.8	1.1	1
May 5 until 0400	326	4800	—	0.5	0.6	0.5
May 29 after 0200	328	4600	6400	1.0	1.0	1
May 30 until 1200	328	4300	7150	0.3	0.3	0.5
May 10 0500–2000	—	5300	7600	0.4	0.1	0.5
Average for undisturbed days	303	6050	8960	0.2	0.2	0

[1] American magnetic character figure, based on observations of seven observatories.
[2] An estimate of the severity of the ionosphere storm at Washington on an arbitrary scale 0, ½, 1, 1½, and 2, the character 2 representing the most severe disturbance.
* This value was obtained by adding 800 kilocycles to $f_{F_2}{}^o$. See text.

absorption as usual during ionosphere storms was so high that the x component of the F_2-layer reflections could not be observed for several hours during the middle of the day, with the transmitter power and receiver sensitivity available. On the afternoon of May 11 the intensities of sky waves at broadcast frequencies, reflected from a low layer (D layer), rose to about 3 times normal and at night, when reflected by E layer, fell to about 1/10 normal. The weak night sky-wave intensities persisted for several nights after other indications of the storm had passed.

The ionosphere storms which occurred in May are shown in Table I approximately in the order of their severity.

Table II shows the number of hours the night $f_F{}^x$ and daytime $f_{F_2}{}^x$ differed from the average for the undisturbed days of May by more than the given percentages. The disturbed hours are those listed in Table I.

ic E layer seem to account for irregular transmissions at ultra-high frequencies of 56 megacycles or higher. The distances involved are usually those of one-hop E-layer reflections not exceeding about 1600 kilometers.

TABLE IV

Sporadic-E Frequencies in Megacycles

Midnight to Noon
354 Hours of Observation

	Hour, E.S.T.											
Date	00	01	02	03	04	05	06	07	08	09	10	11
9							4.5	8	8	8		
10	8		8		4.5							
19	4.5			4.5	4.5		8	8				
22			8	6	8		8	6				
27							8	8	4	8	8	8
28	4.5										6	8

Noon to Midnight
344 Hours of Observation

	Hour, E.S.T.											
Date	12	13	14	15	16	17	18	19	20	21	22	23
9						4.5		8				8
11					8	8	8	4.5				
18						6	8	4.5	8			
20							6	8	6	4.5	4.5	
22			4.5				4.5			8	6	
26			8			4.5	8	4.5	4.5			
27	8	4.5						4.5		8	4.5	4.5
28						8	8					

Table IV gives the approximate upper limit of frequency of the stronger sporadic-E reflections at vertical incidence. It shows the hours of the day for the days of May, 1938, during which strong sporadic-E reflections were most prevalent at Washington. This table indicates the degree of prevalence of sporadic E layer in the latitude of Washington. It does not indicate the specific times of occurrence at other locations even in the same latitude. Approximate values for maximum usable frequencies for sporadic-E transmission with the reflection point over Washington may be obtained by multiplying the figures in the table by the following factors for the given distances: 500 kilometers, 2.2; 1000 kilometers, 3.6; 1500 kilometers, 4.6.

BOOKLETS, CATALOGS, AND PAMPHLETS RECEIVED

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1-kw Broadcast Transmitter • • • Collins 20H and 20J; specifications and general description. (4 pages, 8½×11 inches, printed.)—*Collins Radio Company, Cedar Rapids, Iowa.*

TUBES

Tube Data (national union) • • • Electrical and mechanical data on 5 phototubes, each available in 2 sensitivities. (2 pages, 8½×11 inches, printed.)—*National Union Radio Corporation, 57 State Street, Newark, N. J.*

Tube Data (rca) • • • The following "Application Notes" have been received: No. 93, inverse-feedback circuit for resistance-coupled amplifiers (7 pages); No. 94, operation of the 6AF6-G (7 pages). (8½×11 inches, printed and lithographed.)—*RCA Manufacturing Company, Inc., Harrison, N. J.*

Tube Data (raytheon) • • • "Engineering Service" technical data sheets have been received on the following new tubes. For use in battery-operated receivers: 4A6G, twin-triode output tube; 1C5G and 1A5G, pentode power amplifiers; 1A7G, heptode for frequency-converter service; 1H5G, combination diode and triode; and 1N5G pentode high-frequency amplifier. For 6.3 volt service: 6AD6G and 6AF6G, indicator tubes, and 6AE6G, control tube for use with them; 6K8G, triode and hexode oscillator-mixer; 6P5G, triode amplifier; and 6S7, pentode amplifier. For remote control of line-operated devices: 0A4G, cold-cathode triode.—*Raytheon Production Corporation, 55 Chapel Street, Newton, Mass.*

Tube Data (westinghouse) • • • Information Bulletin No. 16 describes a new thyratron, WL-629. (4 pages, 8½×11 inches, lithographed.)—*Westinghouse Electric and Manufacturing Company, Bloomfield, N. J.*

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